PROCEEDINGS

37TH ANNUAL BROADCAST ENGINEERING CONFERENCE

NATIONAL ASSOCIATION OF BROADCASTERS
LAS VEGAS, NEVADA
1983

PROCEEDINGS

37TH ANNUAL BROADCAST ENGINEERING CONFERENCE

NATIONAL ASSOCIATION OF BROADCASTERS
LAS VEGAS, NEVADA
1983

These proceedings contain technical papers presented at the NAB Engineering Conference April 12 - 13, 1983, in Las Vegas, Nevada.

Published by the NAB Office of Science and Technology

References to the papers contained in the *1983 Proceedings* may be made without specific permission but attributed to the 37th Annual NAB Broadcast Engineering Conference.

Additional copies of the *1983 Proceedings* are available to NAB members for \$20.00 each and to non-members for \$60.00 each.
Please send your request and payment to NAB Services, 1771 N Street, N.W., Washington, D.C. 20036.

NATIONAL ASSOCIATION OF BROADCASTERS
1771 N STREET, N.W., WASHINGTON, D.C. 20036 · (202) 293-3500

THOMAS B. KELLER
SENIOR VICE PRESIDENT
SCIENCE AND TECHNOLOGY
(202) 293-3557

Dear Reader:

For the first time in the history of NAB Annual Conferences, Proceedings are being published *ahead* of the 1983 Conference in Las Vegas for your use during and after the Conference.

These Proceedings present the 1983 technical state of the industry. Within this document are papers on important technical issues concerning all broadcasters (spectrum management and satellites), timely papers on new uses of our broadcasting spectrum (FM-SCA, digital audio and advanced television systems), reference papers (technical descriptions of five AM stereo systems), 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 implement the new ideas and technology presented here. In many respects they are blueprints for our future; a how-to manual of success, and a useful reference manual to compliment the NAB Engineering Handbook.

We at NAB are proud to publish these Proceedings. Your comments on any of the papers or any aspect of the '83 Convention are always welcome.

Best personal regards,

Thomas B. Keller

Thomas B. Keller

1983 BROADCAST ENGINEERING CONFERENCE PROCEEDINGS

CONTENTS

UHF TRANSMISSION SYSTEMS

SATELLITE SYSTEMS

TRANSMITTERS AND ANTENNAS

AM-FM ALLOCATIONS AND FM SCA

TELEVISION AUDIO

NON-IONIZING RADIATION

ADVANCED TELEVISION SYSTEMS

RADIO NEW TECHNOLOGY

AM STEREO: USERS REPORTS

SPECTRUM MANAGEMENT

Reducing Operating Costs and Improving

Performance in Older UHF Transmitters

David C. Danielsons

Harris Corporation

Quincy, Il.

The cost of operating a UHF broadcast station today has risen sharply over the past several years. Electrical companies have been granted large rate increases, some of which are on the order of 30% to 50%. This means that for a 55kW station, running 16 hours per day, the electrical bill can exceed $5500 per month at $.06 per kWhr. These rising operating costs have forced the broadcaster to look for ways of improving the efficiency of the transmitter.

Today's technology now provides the means for the broadcaster to reduce his operating cost, while at the same time improving the performance of his transmitter. Three methods now exist by which this may be done.

The klystrons shipped today by reputable manufacturers will usualy exceed their efficiency specifications because the specified efficiency always represents the minumum performance a manufacturer is willing to ship. Also to insure low phase distortion and good linearity the klystron is never operated near saturation.

By operating the klystron closer to saturation one can gain higher efficiency, at the expense of increased distortion. Figure 1 shows a klystron operating near saturation and the resulting increase of distortion. There is of course a trade off between efficiency and distortion and this is where new technology comes in to play.

New exciters are available which far exceed their predecessor in correction capability. These exciters can easily correct the distortions introduced by the klystron when it is operated near saturation. Todays exciters are capable of correcting up to 20% of linearity distortion and 10 degrees of phase distortion, the

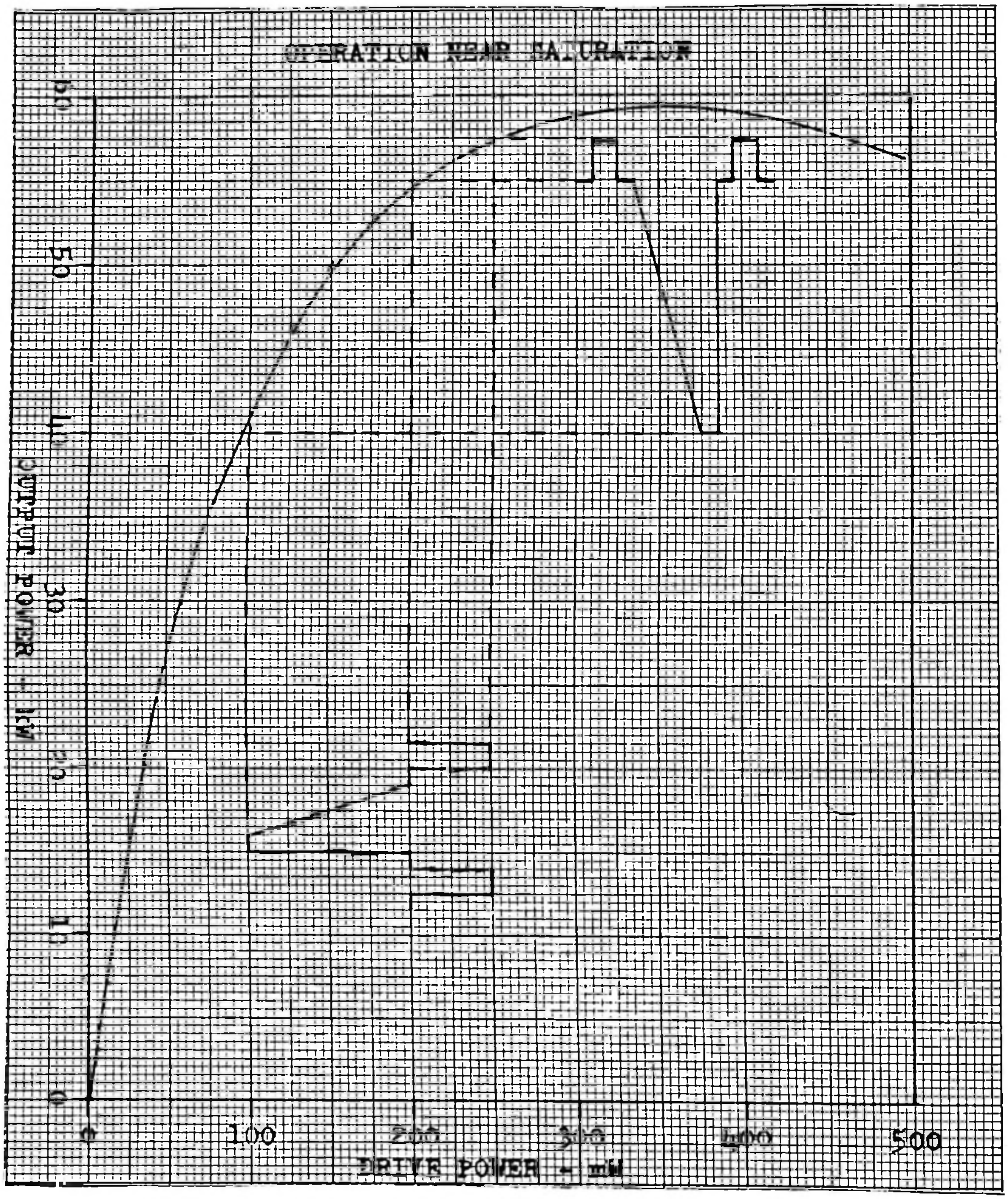


Figure 1

Operation of a klystron near saturation.
Note the sync compression.

two most predominant nonlinearities of the klystron.

Thus, by careful adjustment of his equipment, a skilled broadcast engineer can often increase the efficiency of his transmitter with out sacrificing picture quality.

Combining the mod anode pulser with the new output loading and tuning has shown that beam power can be reduced by up to 35%. This reduction in beam power cuts the operating costs of a UHF station significantly. For a 55kW station this implies a savings of $1400 per month.

In a klystron the beam is velocity modulated by the input cavity and after some distance a componant of the current appears as an amplitude modulation of the beam density. The intermediate cavities enhance this modulation so that at the output cavity the electrons are forced into tight bunches. At the output cavity the electrons are slowed down or stopped as their velocity is transformed into power by the output coupling presenting a resistance to the density amplitude modulated componant of the beam.

When the current is reduced, less electrons are available to be modulated and this causes a reduction in the density amplitude modulated component of the beam. In order to recover the maximum amount of energy from the bunches, a higher resistance must be presented to the beam at the output gap.

By placing a matching network between the output coupling and the rf transmission line, the resistance presented to the beam can be changed. A shorted stub, located 3/8 of a wavelength from the output loop of the klystron , can change the resistance seen by the gap without presenting any reactive compinent which would change the resonant frequency of the output cavity. Figure 2 shows how this occurs on a Smith Chart. If the length of the stub is 1/4 wavelength, then no susceptance is presented to the transmission line. When the stub is lengthened, it presents a capacitive susceptance to the transmission line. The 3/8 wavelength line between the stub and the output loop transforms the susceptance into pure conductance.

A matching network of this type was constructed by Varian Associates, Inc. It consists of a section of transmission line 3/8 wavelength long and a movable shorted stub. Figure 3 shows the basic construction of the coupler.

The coupler was tested in a Harris TV-55U color transmitter operating on channel 26 with an output power of 55kW. The performance of the transmitter was measured without the coupler and without the pulser. Table 1 shows the transmitter performance in this configuration. Figure 4 shows the transmitter tuning.

The coupler was swept on a bench and the stub was tuned for 1/4 wavelength. The coupler was installed in the transmitter and no change in performance was seen.

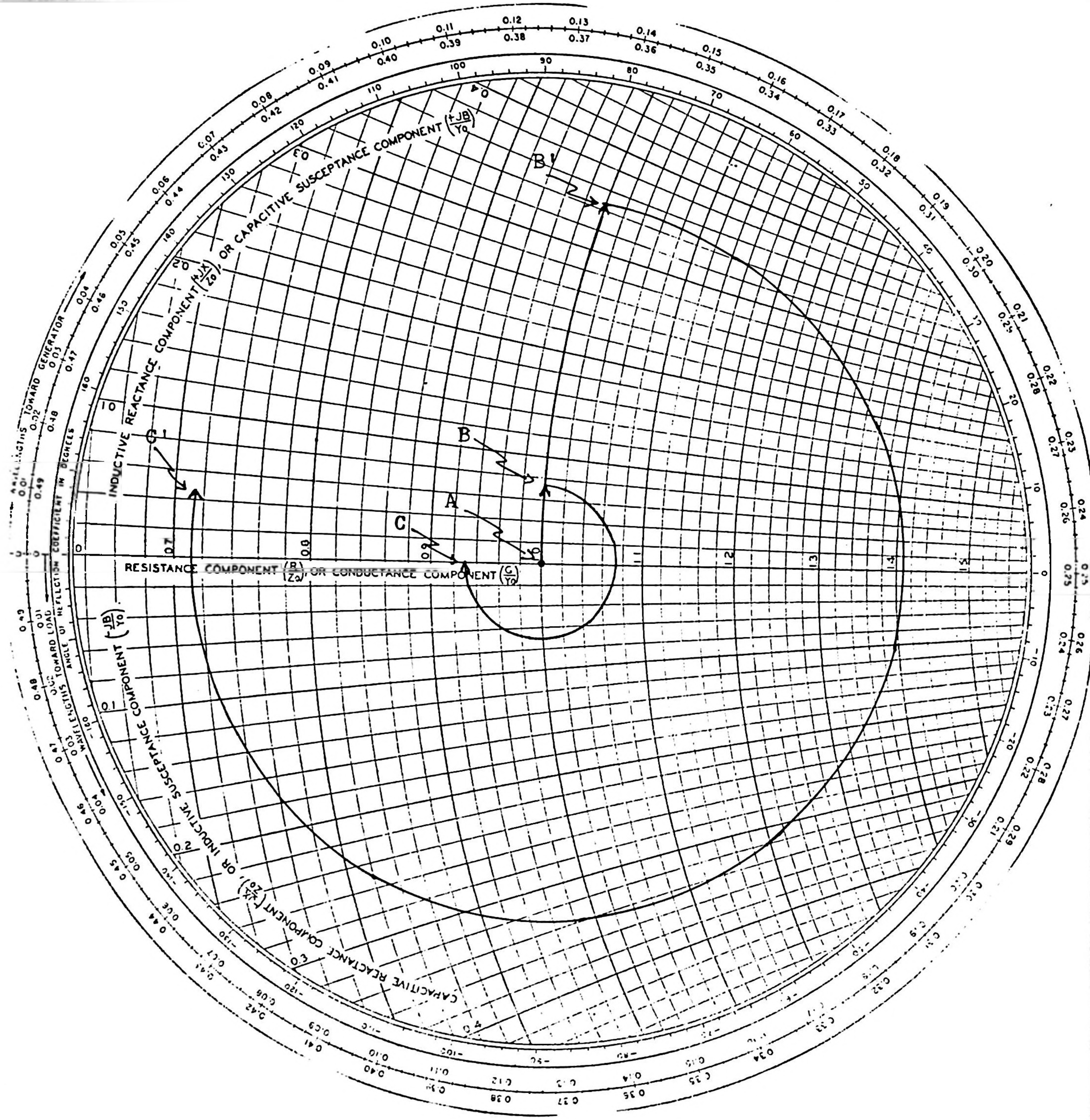


Figure 2

Increased loading of the klystron. Point A represents the 50 ohm line; B- susceptance added by stub; C- susceptance transformed to conductance by 3/8 wavelength line.

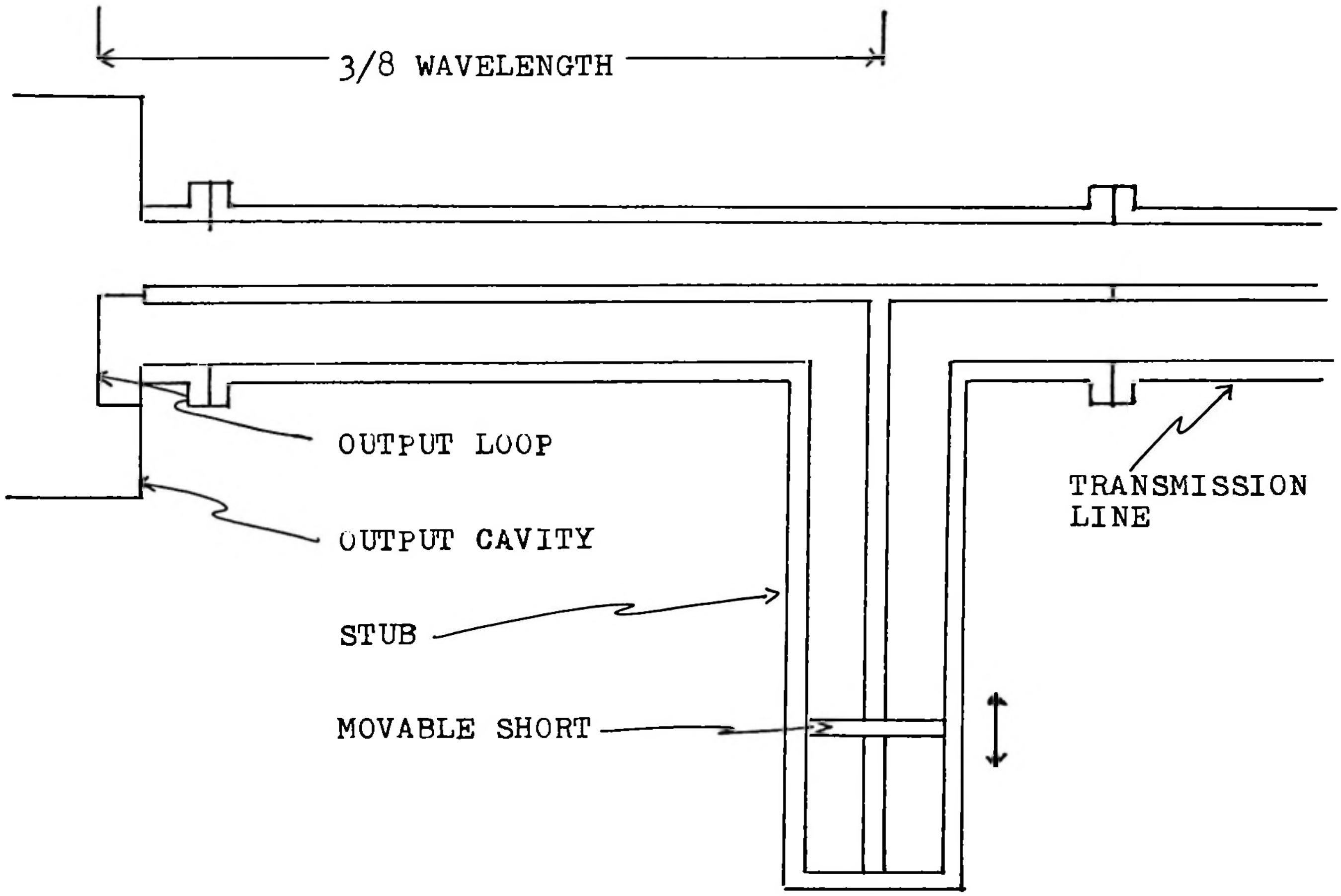


Figure 3

Elements of a coupler.

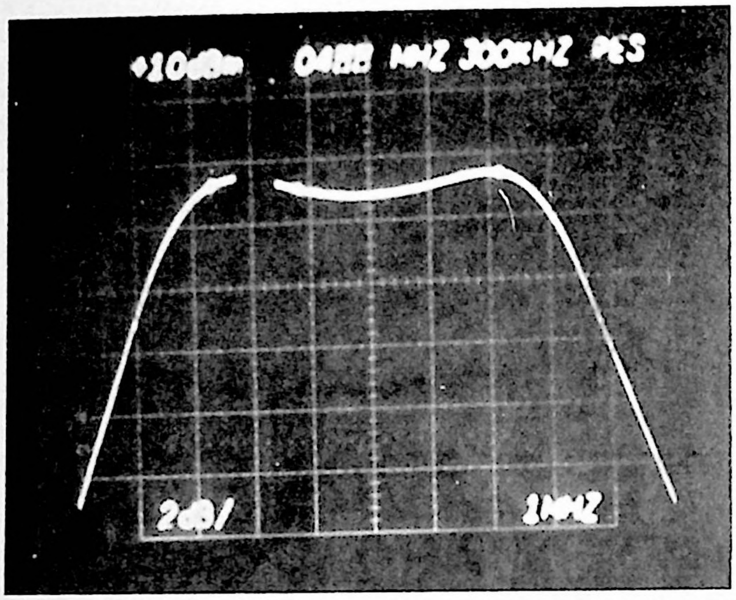


Figure 4

Original transmitter tuning.

The length of the stub was increased by a small amount and an increase in output power was seen and a increase in sync overshoot and sync ringing was seen. By widening the klystron's bandwidth with cavity's 3 and 4, and tuning cavity's 1,2,5, for a flat passband, the increase in sync overshoot and ringing, caused by the raise in the resistance presented to the beam, was canceled.

Beam current was then reduced to lower the output power to 55kW. Cavity's 1,2,5, were again adjusted for a flat passband to compensate for the change in tuning caused by the reduction in beam current.

This process was repeated until a peake in efficiency was found. Final cavity tuning is shown in figure 5. With this tuning and cavity loading beam input power was reduced by 28kW and peak of sync efficiency was raised to 48%.

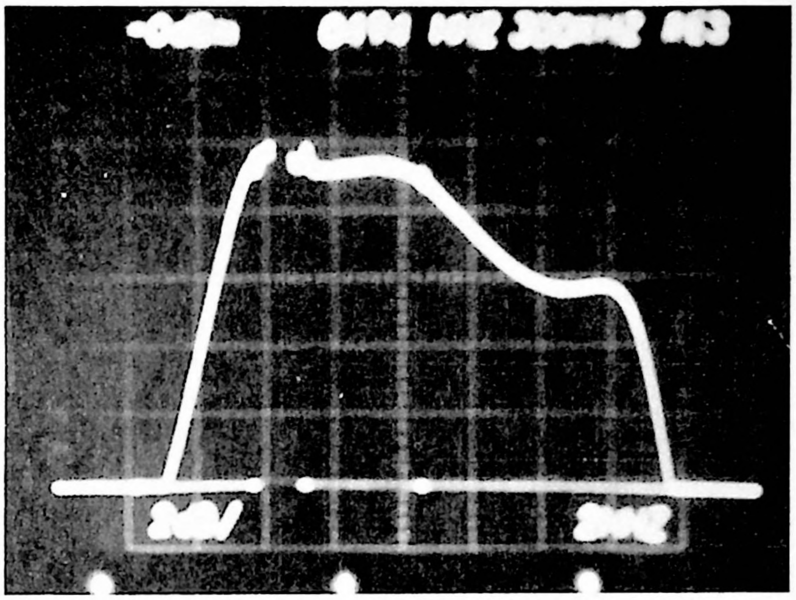


Figure 5

Final klystron tuning with variable visual coupler.

The pulser was then turned on and adjusted, beam input power was reduced by 28kW and peak of sync efficiency raised to 62%. Table 1 compares the transmitters performance before and after. Figures 6 thru 10 are actual photographs of the transmitters response at 62% peak of sync efficiency.

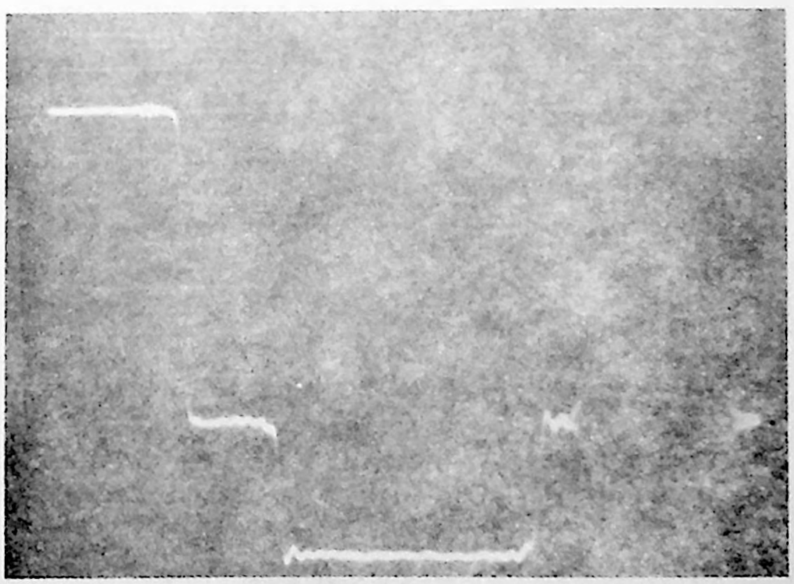


Figure 6
Horizontal interval shows sync overshoot of less than 2%.

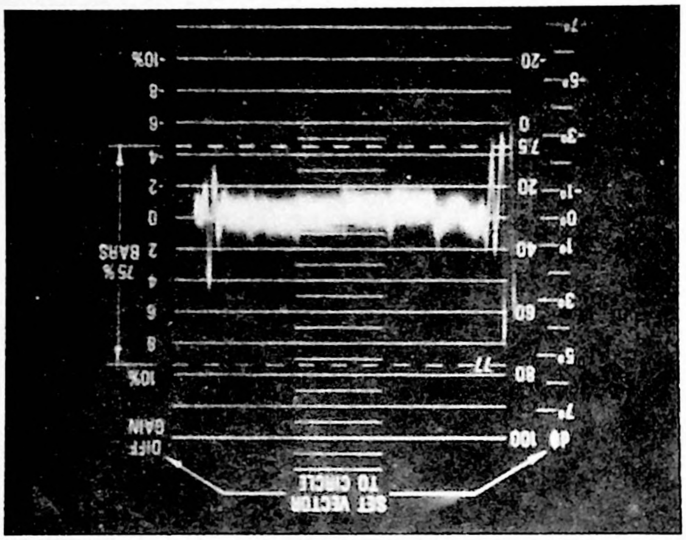


Figure 7

Differential gain is less than 2%

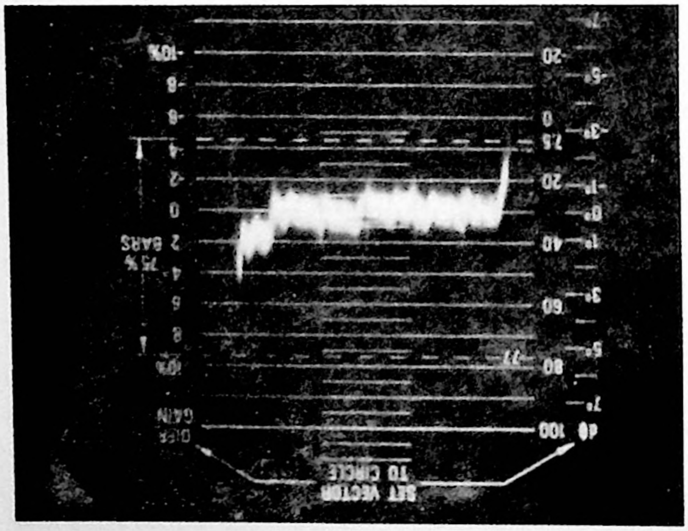


Figure 8

Differential phase is less than 1^{o}.

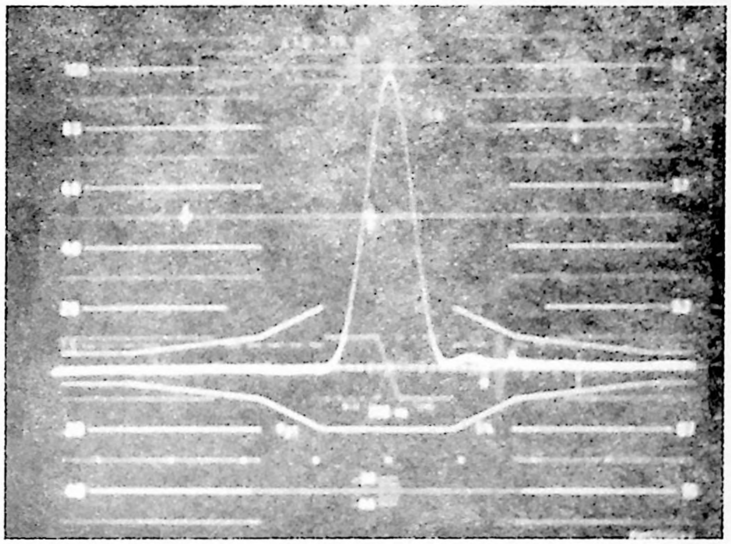


Figure 9

2T Pulse response shows a 1% K Factor.

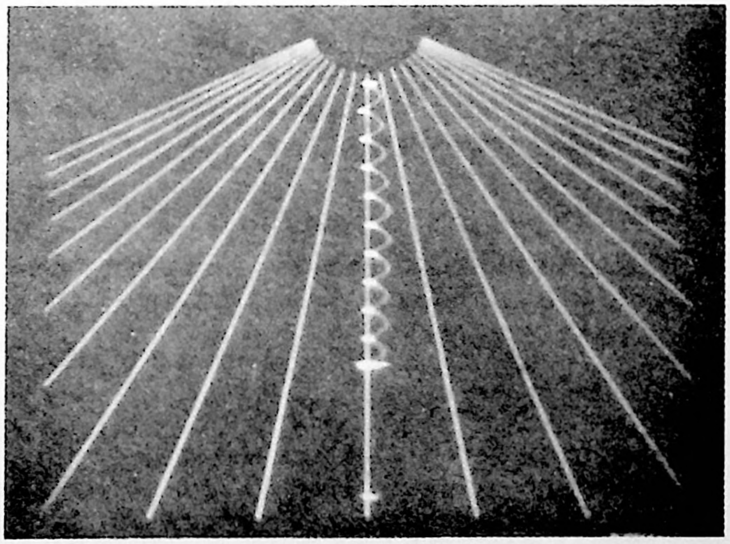


Figure 10

Corrected incidental phase is kess than 0.5°.

Table 1

Parameter	No Coupler No Pulser	Coupler Pulser
Output Power	55kW	55kW
Beam Voltage	24.1kV	24.1kV
Beam Current	6A	3.65A
Beam Efficiency*	38%	62%
Diff Gain		2%
Diff Phase		0.5°
Incidental Phase		$-.5^{\circ}$
2T K Factor		1%
Delay 12.5T		10nS
Low Freq. Lin.		6%
Reg. Of Output		1%
Variation Of Output		.9%

*Peak of sync

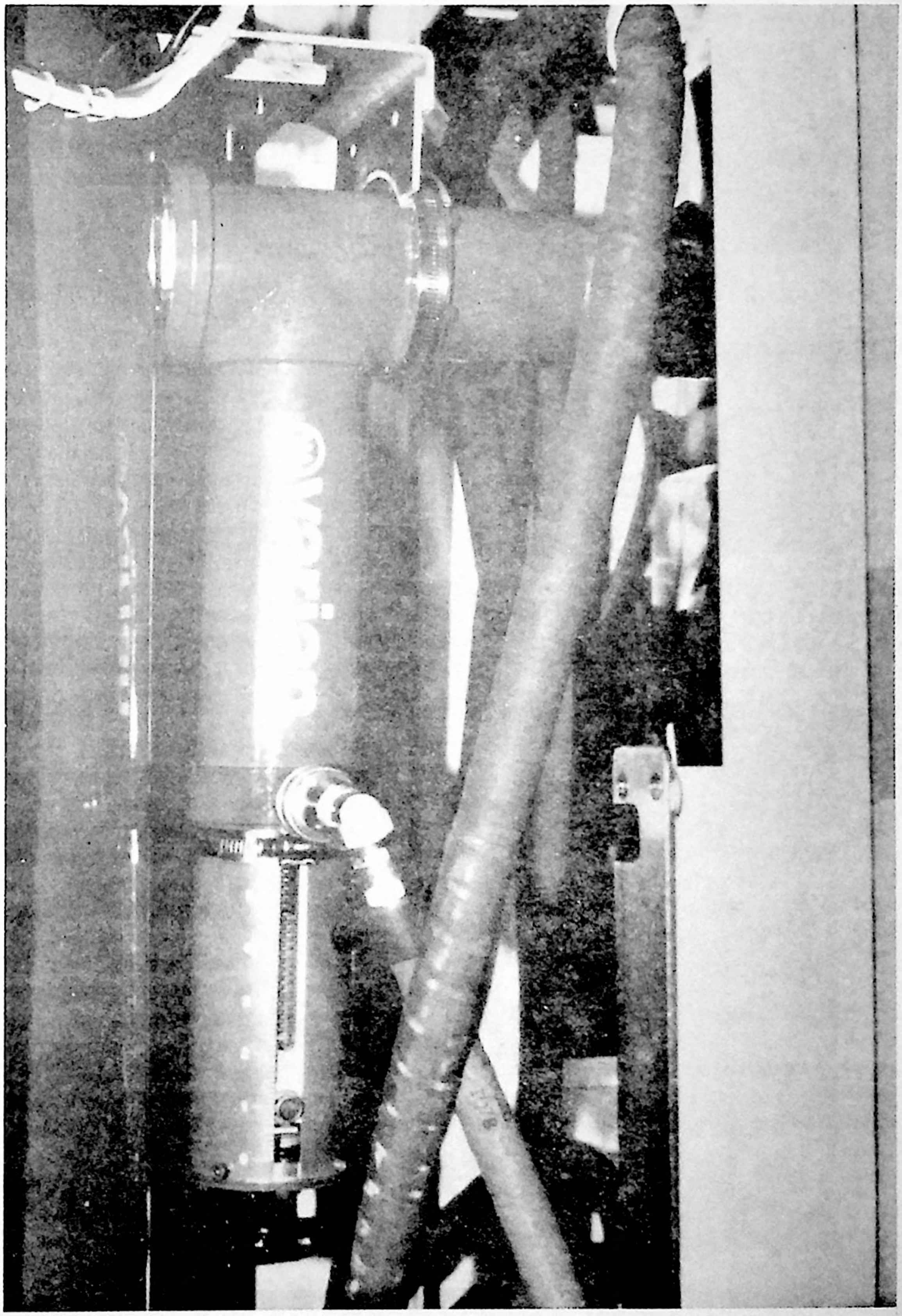


Figure 11

A side view of the klystron shows the variable load coupler in a Harris TV-55U transmitter.

During the previous year several UHF broadcasters have retrofitted their transmitters for efficiency improvements. Among these retrofits are the installation of pulsers in RCA, G.E., and Harris transmitters. Power savings have been substantial, for example WETK, Burlington, Vt. reduced power input by 31% on their RCA-TTU30A.

Variable visual couplers have also been retrofitted in Harris transmitters and are soon to be installed in RCA and G.E. transmitters. Power savings here are also high, WHRO, Norfolk, Va. saved 18.5% by adding a coupler to their BT55U.

The largest savings have accured by adding both pulsers and visual couplers. WUHF, Rochester, NY. saved 38.8% by retrofitting his TV55U.

Table 2 shows a typical list of several UHF broadcasters who have retrofitted their transmitters with pulsers or couplers or both.

Pulsers have been relatively easy to install in UHF transmitters. In a Harris transmitter the pulser mounts directly to the rear of the visual cabinet, see Figure 11. For the RCA transmitter, pulsers were also mounted on the rear of the visual cabinet but on a hinge so that access to the klystron and magnet was available, see figure 12 and 13. The resistor divider network was mounted on a aluminum panel and suspended from the ceiling in the H.V. cage. Figure 14 shows the pulser mounted to a G.E. 60kW.

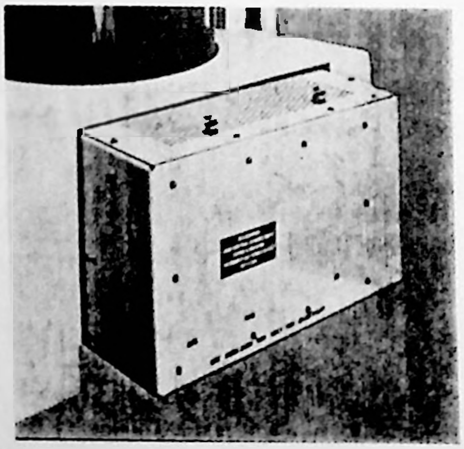


Figure 11

Mod anode pulser mounted on rear of a Harris Uhf transmitter.

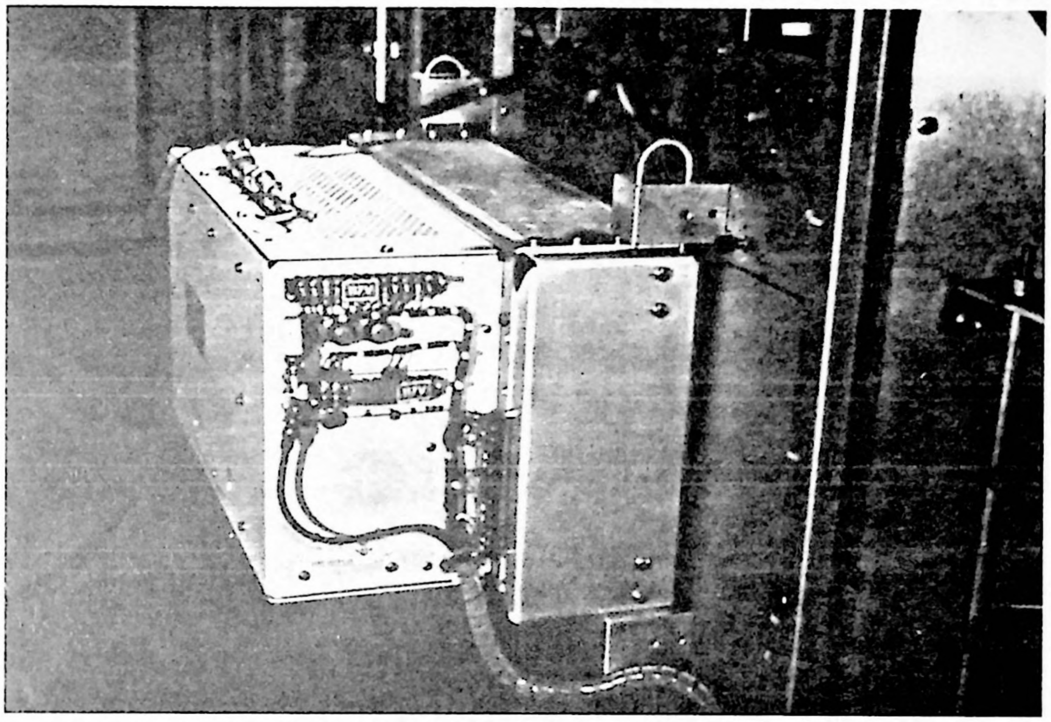


Figure 12

Pulser mounted on the rear of the visual cabinet on WVTA-TV's RCA-TTU30A.

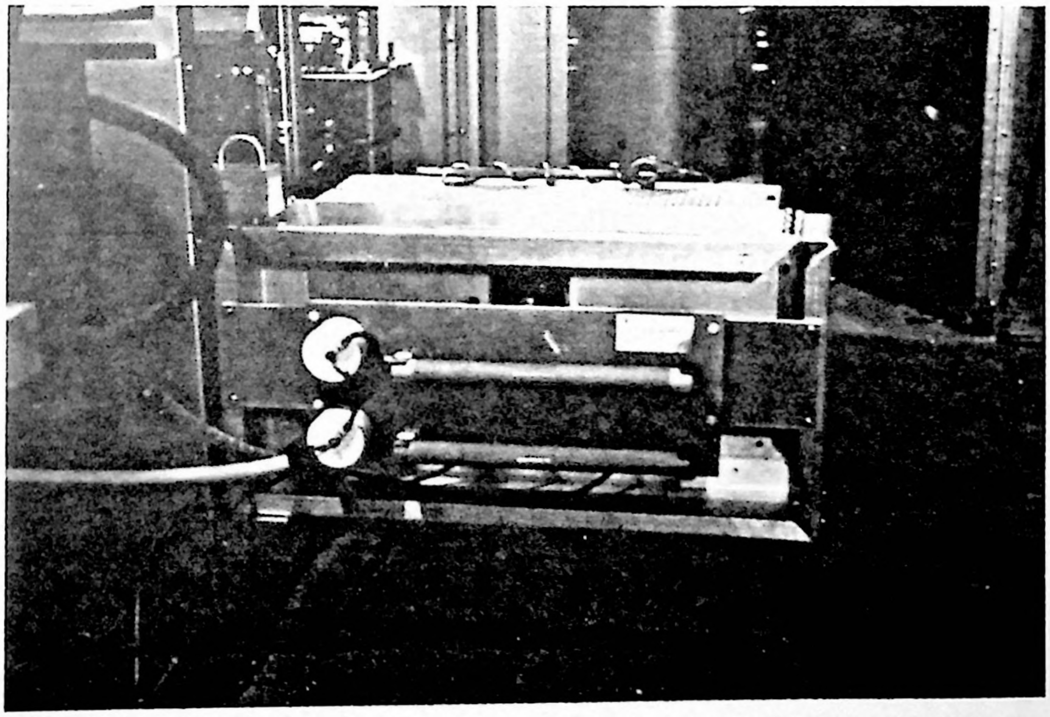


Figure 13

Pulser hinged out to gain access to klystron and magnet.

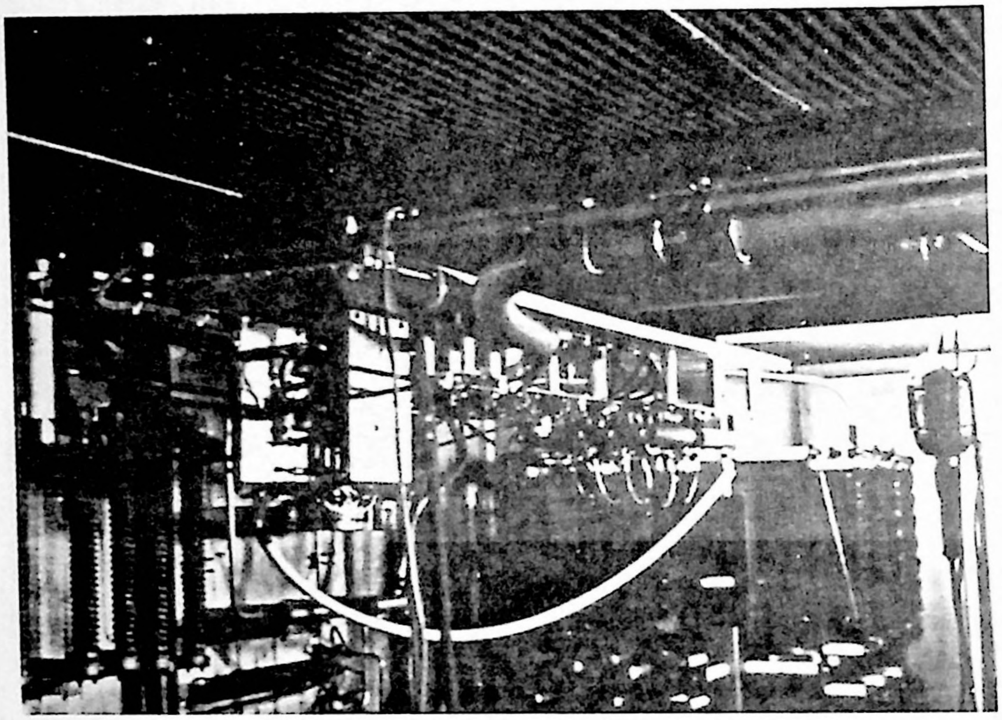


Figure 14

Aluminum panel with zener board, resistor divider, motorized pot. Panel is mounted in H.V. area.

Figure 15

Pulser mounted overhead on a G.E. 60kW at WEDW-TV, Bridgeport, Connecticut.

UHF RETROFIT EFFICIENCY IMPROVEMENTS

					BEFORE					AFTER				
CALL LETTERS	CITY, STATE	CH	XMTR TYPE	TUBE TYPE	BEAM I (A)	BEAM V (KV)	POWER CONS (KW)	EFF %	MOD TYPE	BEAM I	POWER CONS (KW)	POWER SAVING (KW)	POWER SAVING %	EFF %
WVTA	Windsor, VT	41	TTU30A	VA890H	4.7	19.0	89.3	33.6	P	3.3	62.7	26.6	29.8	47.8
WETK	Burlington, VT	33	TTU30A	VA890H	4.85	19.0	92.2	32.6	P	3.35	63.7	28.5	31.0	47.1
WHRO	Norfolk, VA	15	BT55U	VA953G	6.5	25.0	162.5	33.9	C	5.3	132.0	30.0	18.5	41.5
WSNL	Long Island, NY	67	BT55U	VA955G	6.8	24.0	163.2	33.7	P	5.0	120.0	43.2	26.5	45.8
WFLX	West Palm Beach, FL	29	TV55U	VA953H	6.4	24.0	153.6	35.8	P	5.3	127.2	26.4	17.2	43.2
WFLX	West Palm Beach, FL	29	TV55U	VA953H	6.4	24.0	153.6	35.8	C	5.4	129.6	24.0	15.6	42.4
WFLX	West Palm Beach, FL	29	TV55U	VA953H	6.4	24.0	153.6	35.8	P/C	4.65	111.6	42.0	27.3	49.2
WAWS	Jacksonville, FL	30	TV55U	VA954H	6.4	24.0	153.6	35.8	C	5.5	132.0	21.6	14.1	41.6
WAWS	Jacksonville, FL	30	TV55U	VA954H	6.4	24.0	153.6	35.8	P/C	4.2	100.8	52.8	34.4	54.6
WUHF	Rochester, NY	31	TV55U	VA954H	6.7	24.0	160.8	34.2	P	5.5	132.0	28.8	17.9	41.6
WUHF	Rochester, NY	31	TV55U	VA954H	6.7	24.0	160.8	34.2	P/C	4.1	98.4	62.4	38.8	55.9

MODIFICATION CODES: P=PULSER C=COUPLER P/C=PULSER + COUPLER

UHF TV TRANSMITTER PLANT ECONOMIES

John H. Battison

Director of Engineering

WOSU Stations

Columbus, Ohio

Most of the educational/non-commercial television stations operate in the UHF band. At one time the UHF transmitter operator had a very tough row to hoe, not only because of alleged inferior reception but also because of the horrendous cost involved in generating megawatts of power in an effort to compete equally with his VHF brethern. The UHF's burden falls twice as heavily on the educational/non-commercial operator. We have to contend not only with greater viewer resistance and dimishing state or independent operating budgets, but also decreasing support from the various parent university licensees. For these, and other reasons, it has become even more essential for the non-commercial operator to practice every possible economy in these days of exorbitant power costs.

As UHF television transmitter design has continued to improve, tremendous improvements in reduction of power consumption and improvement of efficiency figures have been promulgated. If one believes everything that we hear it is now possible to obtain as much as 60% efficiency in the UHF television transmitter! That is wonderful if you happen to have a new "state of the art" transmitter. Unfortunately, many of the educational stations are the possessors of old, very far from state of the art equipment. WOSU-TV, and WPBO-TV our Portsmouth satellite both use old GE transmitters which were designed about 15 years ago. WOSU-TV in Columbus, Ohio has a 60 KW transmitter and WPBO-TV has the 30 KW version which sports only one Klystron in the visual stage.

The operators of these old transmitters remind me of owners of an older house where heat is lost through inadequate insulation, and leakage through walls and windows. It is surprising what economies can be effected around the transmitter plant itself, even before one pays attention to the transmitter, by shutting off leaks of a few kilowatts here and there.

At WOSU-TV I discovered that the transmitter building did not have a main disconnect switch! The specifications that I unearthed showed that a main disconnect switch had been called for, but somehow it had not been installed, and the state electrical inspectors had failed to note its absence! This in itself may seem a small thing. However, it involved us in about $2500.00 in repair and overtime costs, as well as a very uncomfortable 12 hours in an unheated building working inside and out in many inches of snow replacing a burned out primary feed. Because there was no disconnect, the main breaker had been overworked as a means of disconnecting the transmitter at the close of each operating day. Over the years this improper use had worn the breaker so that one day -- naturally it was below zero and had been snowing hard -- it caught fire. The resulting fire burned the cable insulation, ruined the breaker, the three phase metering transformers, and in general caused a lot of problems on a below zero Saturday afternoon and evening! As I shall relate later in this paper this lack of disconnect had been costing us wasted power for many years.

If you mention the word "economy" and UHF transmitter operation in the same sentence most engineers will immediately think of improvements in transmitter efficiency. This is natural, because by their training, engineers are predisposed to look toward efficient operation. The use of hundreds of kilowatts of power with a transmitter output of 25 to 60 KW certainly galls the economy/efficiency minded engineer. His immediate reaction is to start looking for ways of improving transmitter circuitry and operating parameters in efforts to raise the operating efficiency above 30%!

Unfortunately, owners of older and more inefficient transmitters are frequently circumscribed in their choice of improvements. Therefore after doing as much as possible to improve the transmitter the only other place to look is in the operation of the transmitter plant itself. It is here that surprising reductions in total power consumption can be made by careful use of load control philosophies. The normal method of measuring power consumption as used by the power companies is to install a demand meter and the usual consumption meter. Methods of power consumption and charges therefore, vary in different parts of the country and in even different parts of the state! We operate WOSU-TV in Columbus, Ohio and WPBO-TV in Portsmouth, Ohio. Each transmitter is supplied by a different electric power company. Each company has a different method of calculating power costs. But each method has something in common. The higher the demand meter reading, the higher the overall bill.

In Portsmouth every kilowatt of demand costs us approximately $5.00. In Columbus the demand after the first 50 kilowatts, which cost $520, is approximately $5.00 per kilowatt hour. This demand meter reading is also used in calculating the total power consumed, for billing purposes. The power company has explained to me several times the method of applying the demand figure to the calculations. It is a little complicated and certainly quite time consuming, so I shall not go into it here. The important thing to remember is to keep the demand meter reading as low as possible. Every kilowatt that meter reads will cost another $5.00, or more, depending on your local supplier.

I mentioned differences in applying the demand meter reading to power consumption calculations. In the Columbus area the demand meter records the highest figure used in a half-hour period. At Portsmouth, the demand meter records the highest figure used during a fifteen minute period. These figures are progressive in that if the highest demand recorded in a given month is 150 kilowatts and a heater load of, say 20 KW, comes on for 35 minutes you will be charged an additional 20 KW at $5.00 per kilowatt hour, or $100.00. This is in addition to the consumption recorded on your regular meter. Thus 35 minutes use of an electric heater has increased your month's power bill by at least $100.00! Pretty expensive comfort isn't it? Now in the case of our Portsmouth transmitter it would have taken only 20 minutes to hit the $100.00 mark because of the shorter, 15 minute, demand meter measuring period.

The main thing that becomes apparent is to keep demand figures as low as possible, and avoid putting additional power consuming equipment on circuit when the transmitter is in use. When the transmitter is not operating, any loads totalling less than the transmitter's normal consumption will not affect the demand meter. Similiarly, once a given reading has been reached on the demand meter a combination of loads totalling <u>less</u> than this recorded figure will not change its reading.

Our transmitters at Columbus and Portsmouth are both manned; therefore human comfort must be considered in addition to environmental requirements for the transmitters. Normally there is sufficient heat from the transmitter in the winter to keep the operating areas at a comfortable temperature; thus additional heat is very seldom required in the daytime. At night, between sign-off at midnight and sign-on at 6 a.m., electric heaters (50 KW) come on as required to maintain a reasonable temperature and to prevent freezing of the transmitter coolant. I found that in the past these heaters were frequently switched on while the transmitter was still operating. This resulted in a load overlap so that our demand would go up by 50 KW! Careful observation by the transmitter crew of load control eliminated this problem.

In the summer, air conditioning is necessary to maintain a reasonable environment. Fortunately it has been possible to control the compressors so that they run for less than half-an-hour at a time in the case of Columbus, thus keeping the recorded demand to a minimum.

In the case of our Portsmouth transmitter, where the demand recording period is only 15 minutes, such short compressor "on" time is usually impractical and in the summer we have to live with the increased load.

Another source of unnecessary power consumption can be found in the heat exchanger system. Our GE transmitter came with two heat exchanger pumps for the 60 KW unit. Much of the year it is possible to run with only one pump load in use. This saves us 9.1 KW per hour. As soon as the Klystron temperature reaches 145° the second pump comes into circuit. Unfortunately this was not possible for our 30 KW transmitter at Portsmouth, because only one pump is used.

Another 7.5 KW leak was discovered at the voltage regulator. Again, because of errors in the original installation over 10 years ago, provision had not been made to disconnect the voltage regulator when the transmitter shut down. Therefore, all through the night when the transmitter is not in use the voltage regulators continued to operate and impose a load of 7.5 KW. Intelligent rewiring and reconfiguration of control circuits eliminated this unwanted load!

It might be interesting to consider for a moment electrical wiring considerations in transmitters. Many of us are not fortunate enough to be in on the ground floor, and able to plan and design our transmitter installations exactly the way that we would like to have them. Usually we inherit somebody elses' "baby", which may or may not be a well disciplined infant! It is very easy to go into an installation and, taking an overall view, continue to act on the assumption that your predecessors were very power cost conscious, and had done everything to ensure the highest efficiency of operation. In our case, it was not until the energy crunch came that we looked into sources of unwanted loads and wasted power. I'm not advocating a penny pinching, skinflint, miserly type of operation in which everybody has to switch off lights the second they leave a room because this can be self defeating. There's an old adage "take care of the pennies and the dollars take care of themselves". Today with energy costs being what they are ,saving the odd two or three hundred watts can add up to a sizeable sum. Merely changing over from incandescent to fluorescent lighting can make a noticeable change in power consumption if the area to be illuminated is large. Of course, the cost of the changeover must be weighed against the power costs saved, and the pay-out period has to be reasonable.

Now, turning to the transmitter itself; what can one do to improve its efficiency? Probably the first thing that comes to mind these days is to add mod pulsing. This is fine -- if you have a reasonably new transmitter that uses IF modulation and is easily adapted to this modification. Another change that can be made is switching from internal, to external, cavity Klystrons. This involves the addition of new magnetics and poses problems of tube compatibility and interchangeability. For example, our 60 KW transmitter uses three Klystrons, one in the audio and two in the visual PA. The question of changing from internal to external has to be considered in light of the tube complement. Another possibility is the installation of output couplers. Again, the feasibility of this modification depends on the type of Klystrons used and the transmitter itself.

Many of the older transmitters employed bias supplies to the modulating anode whose output voltage was controlled by taps. This provided adequate control of voltage so that the transmitter operated properly but, perhaps, not at maximum efficiency. We bought two of the variable power supplies made by George Townsend and immediately were able to save nearly two amps at 18.5 KV through proper adjustment of this voltage. No compromises in terms of band width standards were made, and this resulted in a savings of about 30% of our previous power consumption.

Another device that we added, while not being strictly in the field of power consumption reduction, certainly earns a place for itself in the area of operating costs reduction. From about 1978 on we began to incur multiple power surges, spikes and all kinds of line disturbances on our primary power. This became so bad that we began to lose stacks of 50 rectifiers at a time. Conversations with the power company had no effect. So we installed a set of the late Duffy Wilkinson's surge protection equipment. Whether it was just luck, providence, some cooperation from the power company or just the plain fact that Duffy's surge protector works, we haven't lost a single diode since then! The $1,000.00 that it cost for the surge protection equipment was very well worth it in terms of overtime hours, replacement parts and general morale.

We looked into the use of mod pulsing with our transmitter and for a couple of years it appeared to be completely out of the question because of the need to spend about $100,000.00 for a new driver that would enable us to use it. We also considered the use of two English Electric 55 KW high efficiency Klystrons in place of our two 30 KW Klystrons in the Columbus transmitter. Unfortunately although this would have resulted in considerable operating efficiency, the cost of new magnetics required made the project unrealistic in terms of return on our investment. Even although Harris has now come out with a replacement driver stage at a very reasonable price, the cost of the driver stage, and adding mod pulsing, would still run well over $100,000 per transmitter and I don't feel that we would recoup our investment before the transmitters are scrapped. Therefore we have decided to operate our transmitting plant as efficiently as possible until replacement funds for the transmitters become available. In the case of the Columbus transmitter we have applied for a grant to replace the antenna, transmission line and transmitter. At this point we are concentrating, and have been for the last few years, on obtaining the most efficient operation possible. This has resulted in overall plant efficiency of about 40% which is a great improvement over 25%.

I've not touched on one improvement that is usually beyond the scope of the average station unless they happen to be in a very fortunate situation. We use 6 1/8" coax to the top of our 1000 ft. tower. Everyone knows losses in a line of this size at 600 mHz are not negligible. Part of our new project is the installation of waveguide. Fortunately our tower is stressed to carry this at the size required, and we shall have the choice of maintaining our ERP and reducing operating costs, or increasing ERP and reducing our operating costs to a lessor extent through the power losses that will be eliminated in the waveguide.

A closing point of interest is that Channel 28, WTTE-TV, will be joining us on our tower this summer. In the meantime, a new building is under construction to house their transmitter as well as our new one. We are planning joint use of circular waveguide to serve both WOSU's channel 34 and WTTE's Channel 28 antennas. Fortunately the tower was planned by foresighted people many years ago and it was designed for three antennas and associated transmission lines.

New Developments In High Efficiency

High Power U.H.F. T.V. Klystrons

Dr R Heppinstall

English Electric Valve Company Limited

Chelmsford Essex England

Introduction

The prime interest of the U.H.F. T.V. broadcaster is to transmit a signal which meets the required specification at minimum cost. The cost of ownership of the transmitter itself is therefore of considerable importance to him and a significant contribution to this is the operating cost of the high power amplifier. Klystrons have performed admirably in this service for many years, with a well proven record of long life and reliability. During recent years klystron manufacturers and transmitter manufacturers and users alike have been investigating various improvements in klystron design and operating techniques directed towards achieving an appreciable increase in klystron efficiency with consequent reductions in the overall operating cost of the transmitter. English Electric Valve Co. is currently carrying out a comprehensive development programme in this field and this paper describes some of the results obtained from this work.

Sync. Pulsing

The operating efficiency of a klystron amplifier in a U.H.F. T.V. transmitter can be defined as the ratio of peak sync. output power to average beam power. It can be increased appreciably by adopting the sync. pulsing technique. This consists of pulsing the voltage of a suitable electrode in the electron gun of the klystron so that the beam current and hence the beam power is high during the sync. pulse period and much lower during the picture period. This technique can be applied to three types of electron gun design which can be characterised by the magnitude of the voltage swing required to produce the required change in beam current.

Investigations have been carried out on klystrons having electron guns of the traditional design in which the modulating anode is pulsed. Relatively large voltage swings are required but nevertheless this technique has become established as economically viable, particularly in high power transmitters.

Recently klystrons have been manufactured in which the electron gun incorporates a beam control device (BCD), similar to the focus electrode in the conventional electron gun. When this BCD electrode is connected externally to the cathode the klystron's performance characteristics are the same as that of a standard klystron. However if a negative voltage (with respect to cathode) is applied to the electrode the beam current is reduced. Medium amplitude voltage pulses are required to produce significant beam power savings. Klystrons incorporating beam control devices are now in production.

Another arrangement for producing changes in beam current is to incorporate a shadow grid structure in the electron gun - a pair of carefully matched, carefully aligned grids are mounted close to the cathode surface. EEV has produced samples of its 40kW K3217H klystron incorporating a shadow gridded electron gun (Ref. 1). Experimental investigations, conducted under test conditions at the BBC's Crystal Palace transmitter have provided valuable results on the performance of a pulsed klystron in a U.H.F. T.V. transmitter.

In order to optimise the transmitter efficiency when the klystron is being operated in the beam current pulsed mode it is necessary at all times to have just sufficient klystron beam power available to produce the required R.F. output signal. However there are a number of practical factors which may provide a limit to the efficiency which can be achieved. These factors are the klystron tuning method adopted and the provision of an R.F. drive with the required correction features.

An experimental investigation has been made to establish guidelines on the method of tuning the klystron for pulsed operation. The T.V. waveform is such that the picture period is present for a high proportion (about 93%) of the line period. However tuning the klystron to give optimum performance during the picture period - in order to obtain the best overall efficiency - results in an output cavity loop position which means that the loop is significantly undercoupled during the sync. pulse period. This leads to unsatisfactory performance. It is therefore important that the output cavity loop is positioned so as to provide maximum output power during the sync. pulse period - a condition which corresponds to being overcoupled during the picture region. The klystron should then be tuned using the remaining controls to give the required frequency response for the picture region operating conditions.

Examination of the R.F. transfer characteristics (Figure 1) shows that the R.F. drive power needed to saturate the klystron when operated at the voltage and current corresponding to peak sync. output power is appreciably less than that needed under the picture operating conditions. The driver must therefore meet this requirement. The driver must also have far greater correction facilities than were previously necessary. During the picture region the saturated output power of the klystron will be much closer to black level and therefore the non-linearity of the klystron will be appreciably greater.

The results are shown in Tables 1 and 2.

Beam voltage	(kV)	23.0
Beam current	(A)	2.6
Output power, black level	(kW)	23.5
Input power, black level	(W)	4.2
Input power, peak sync.	(W)	3.0
Grid voltage, black level	(V)	136
Grid voltage, peak sync.	(V)	416
Grid voltage swing	(V)	280
Effective peak sync. efficiency	(%)	70
Basic klystron efficiency	(%)	40

Table 1: Gridded klystron pulsed performance.

		Before correction	After correction
Linearity	%	20	99
Differential Gain	%	47	95
Differential Phase	Deg	5	1
Peak sync. incidental phase	Deg	63	4

Table 2: Gridded klystron : T.V. performance.

These show that an increase in operating efficiency from 40% in the unpulsed mode to 70% in the pulsed mode has been obtained whilst at the same time the T.V. parameters of linearity, differential gain, differential phase and incidental phase have all been corrected. The saturated output power of the klystron during the picture region was 24.9kW, giving a black level to saturation power ratio of 94%. The beam power in the pulsed mode is only 60kW, a saving of 43kW compared with the corresponding beam power of 103kW in the unpulsed mode. This represents a saving of about 20,000 dollars per year in electricity costs.

The possibility of full time modulation, in which the beam current is controlled according to the instantaneous luminance level, is attractive. The associated correction problems need careful evaluation. However, it is worth noting that the voltage swings required on a gridded klystron would be relatively small - a 70% reduction in beam current can be achieved at the 40kW power level by a swing of about 450 volts. This is well within the capability of a solid state pulser.

The pulsed up mode of operation

The underlying assumption on which the pulse work so far described has been based is that the peak sync. R.F. output power level of the transmitter remains unchanged. An intriguing possibility is to consider using the pulsed technique to increase the peak sync. output power level obtained from an existing klystron. This may be achievable by applying a positive voltage pulse to the modulating anode to increase the beam current during the sync. pulse period. Such a technique is feasible only if the klystron design is capable of providing the required performance at the higher power level. In particular the cathode must be capable of providing the increased current, the output cavity must be capable of transmitting both the increased peak sync. and the increased mean R.F. power levels and the focussing characteristics must remain satisfactory.

The cathodes used in U.H.F. T.V. klystrons are often conservatively rated and many are quite capable of providing an increased current during the sync. pulse period. Further, during routine production testing at EEV klystrons are driven c.w. at the saturated power level for some time, demonstrating that the output ceramic can satisfactorily transmit higher mean R.F. power levels than are generally experienced during T.V. transmission. The peak R.F. transmitting capability of an external cavity is also high - EEV have for many years manufactured an external cavity radar klystron having a peak output power of 500 kilowatts at 600MHz.

An investigation into the pulsed up mode has been made, using an EEV klystron type K370, which normally operates at a beam voltage of 12.5kV and a beam current of 2.8A, giving an output power of 11.5kW. Positive voltage pulses of various amplitudes were applied to the modulating anode and the resultant output power determined. Figure 2 shows the percentage increase in output power as the pulse amplitude is increased. Raising the pulse amplitude from zero to 2kV (corresponding to an increase in beam current from 2.8 to 3.5A) resulted in a significant increase in output power. Further increases in pulse amplitude resulted in no further significant increase in output power due to a reduction in klystron efficiency cancelling out any expected increase in output power resulting from the increased beam power. The beam focussing characteristics remained good for the range of pulse voltages employed.

Based on this preliminary work it seems possible that an increase of the order of 15 to 20% in peak sync. level may be expected from a given klystron by adopting the pulse up technique.

A wideband 40/55kW klystron

One important factor in the cost of ownership of a transmitter is the cost of holding spares. For many years three klystrons were customarily required at each power level to obtain full coverage of the U.H.F. T.V. range (470-860MHz) and this had an adverse effect on spares holding costs. EEV have now designed klystron and circuit assemblies so that at power levels up to about 30kW the full frequency range can be covered by a single tube - for example the EEV external cavity K3270 klystron covers the power levels up to 15kW and the K3271 klystron covers the range up to 30kW. Essentially the increased frequency range has been obtained by constructing klystrons using cavity ceramics of smaller diameter together with cavities having a much longer door travel. Associated circuit assemblies occupying significantly less floor space were also designed.

In principal such techniques can be applied to klystrons designed for the higher power levels. This possibility has been investigated. The initial objective was to construct a klystron using smaller diameter ceramics and covering the combined frequency range of the EEV K3276H and K3277H klystrons (470-704MHz). A further design requirement was that the tube would fit inside the existing K3276H or K3277H magnetic frame. Sample klystrons have been designed and constructed. Specially designed cavities have also been constructed and test work has demonstrated that an upper frequency limit of at least 750MHz has been obtained.

Further work in hand is designed to increase the frequency range to around 810MHz and provide a new compact circuit assembly occupying only about 22" by 22" of transmitter floor space.

Reference

1. "A 40/55kW Gridded Klystron for T.V. Transmitters" by R. Heppinstall, A. H. Whiteside and S. J. Bainbridge. Proceedings of the International Broadcast Convention 1982, pp 101-103.

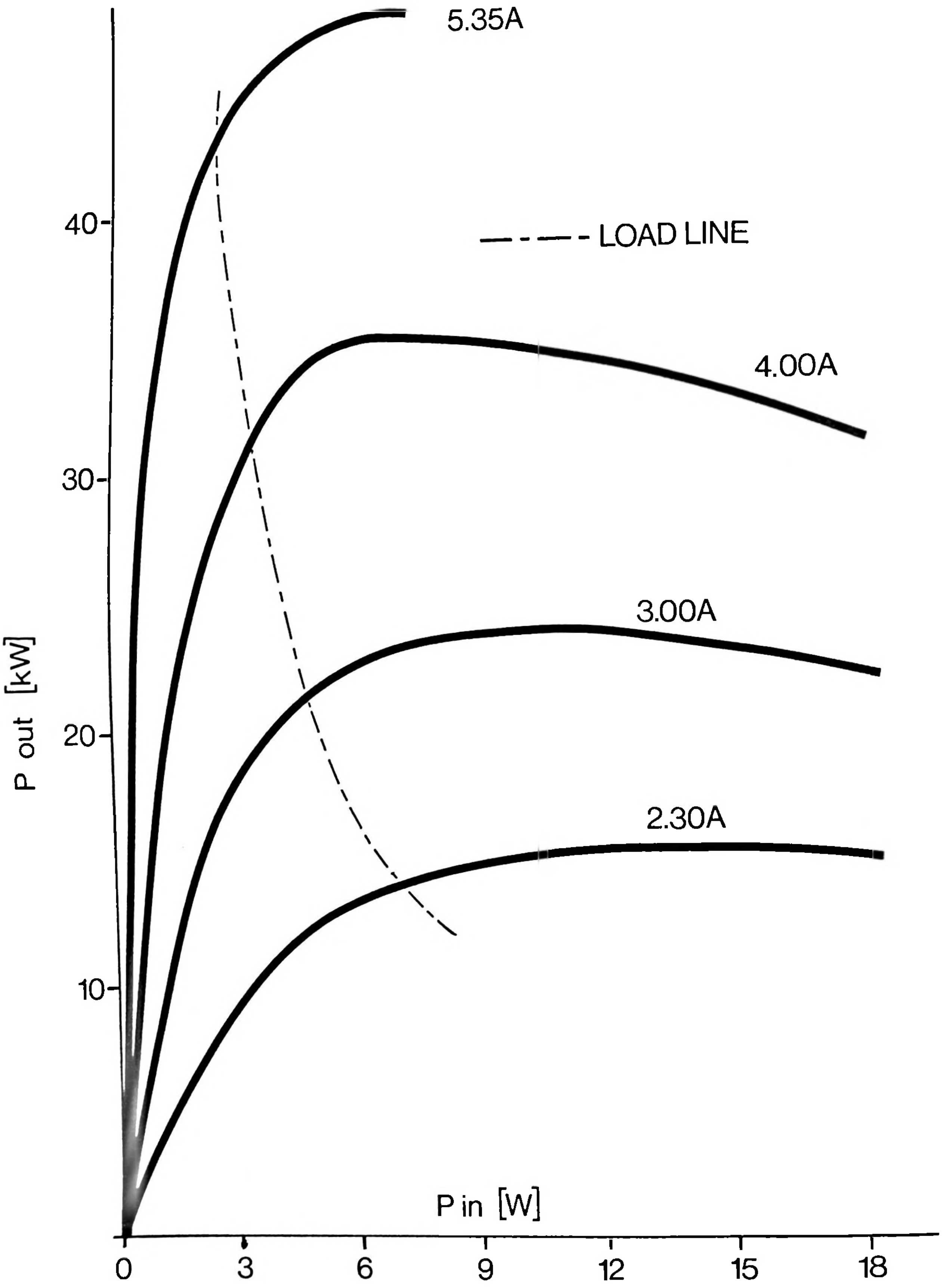


FIG.1 K3217HG—TRANSFER CHARACTERISTICS

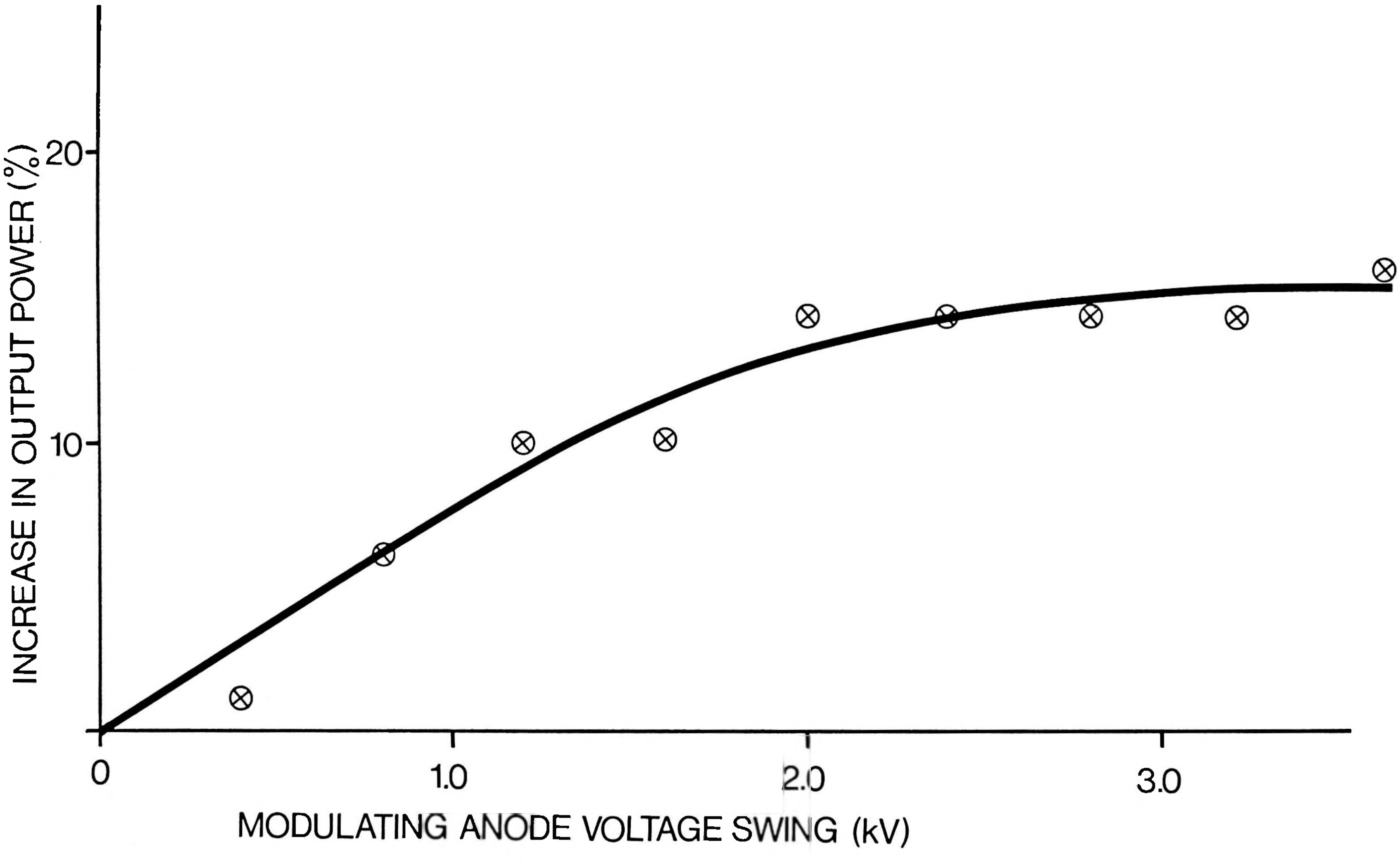


FIG. 2 K370 KLYSTRON — PULSED UP MODE CHARACTERISTICS

INCREASED EFFICIENCY FROM FIVE-CAVITY KLYSTRONS

Robert S. Symons

Varian Associates, Inc.

Palo Alto, California 94303

Recent work at Varian Associates is demonstrating that the efficiency of five-cavity klystrons can be increased significantly by changing the cavity tunings and Q's and increasing drive power. This has been done without making any significant changes in the outline of the tube which would interfere with the use of the tube in existing transmitters.

It is well known that changed tunings and Q's together with changes in drift lengths can lead to higher klystron efficiency[1]. Such tubes will soon become more widely available for UHF television applications. A saturation efficiency of 59% from an experimental 100 kW klystron incorporating such features was recently reported by us[2]. This tube, however, will not be of much help to broadcasters who have recently purchased modern transmitters using our 55 kW klystrons. We have therefore undertaken a parallel program to find out what can be done to improve tubes suitable for use in existing transmitters. Initial results of this program are very encouraging indeed, and it is now safe to say that a great deal can be done.

Most people who have used klystrons are familiar with the fact that inductive tuning (or tuning to a higher frequency) of the cavity just before the output cavity of a klystron will increase the efficiency. This occurs because an electron bunch in a klystron forms around an electron which passes through the gaps as the voltage goes from being decelerating to accelerating, so the speeded electrons catch up with the slowed ones. When this bunch reaches the next cavity, if the cavity is tuned to the drive frequency, and so presents a resistive impedance, the gap voltage will be in phase with the current. As a result, a new bunch formed at this gap will be 90^{o} out of phase with the old bunch as a consequence of the fact that the new bunch will again form about the electron which passes through this gap when this gap voltage is zero. The cure is to tune the later cavities inductively so their gap voltages are out of phase with the electron bunches and the bunching effects of all gaps are cumulative. Figure 1 shows, by means of an Applegate diagram (named after the patent illustrator who drew the figures for the original Varian klystron patent) how electron bunches form in the drift space between klystron cavities and how proper tuning of a cavity creates a gap voltage which has the same phase relation to the bunch as did the voltage that originally formed it. In looking at an Applegate diagram, it helps to realize that the slope of the electron trajectories is proportional to the electron velocity.

From the first uses of multicavity klystrons everyone has tuned the next-to-the-last or penultimate cavity to the high side of the drive frequency. It has not been as generally appreciated that further increases in efficiency could be achieved by tuning virtually all the intermediate cavities to the high side of the operating frequency while driving the tube harder and harder (nor has it been practical because of a lack of drive power). We[3] observed this, first, while working on a 10 megawatt radar klystron which gave 12% bandwidth in 1958. We tried this tuning pattern because theory predicted that it would give the greatest gain - bandwidth product in a short klystron, but we soon found that it also gave the highest efficiency, particularly at the low end of the band.

It turns out that efficiency at the low end of the passband is the only place that efficiency matters in UHF television applications. The sync pulse is essentially a pulse of power at the carrier frequency and only the double sideband portion of the spectrum about the carrier frequency contains significant energy. Voltages of the upper sideband components never drive the klystron to saturation and the gain of a klystron, tuned as has been described, is just as flat, below saturation, as that of any conventionally tuned klystron.

After investigating the described tuning on a computer, which confirmed the efficiency advantage, a special VA-953 was built. In order to obtain a flat passband with the proposed tuning, a much smaller loop than usual was used in the input cavity, and a large loop for connection to an external load was added to the second cavity. The first and last cavities were tuned to the carrier frequency, the second cavity was tuned to the high frequency end of the channel and the third and fourth cavities were tuned well above the passband. Efficiencies of about 55% at saturation were observed between channels 14 and 29 with drive powers below 50 watts and with a variable visual coupler on the output of the tube. This compares with about 50% for a conventional VA-953H, represents a 10% power saving, and indicates that a peak-of-sync efficiency in excess of 65% should be possible in a transmitter using beam current pulsing.

When the third and fourth cavities were brought closer to the band, gain equal to that of a standard VA-953H was achieved with slightly higher efficiency, but the 10% reduction in input power needed the higher drive levels.

Figure 2 shows phase-space diagrams for a conventional VA-953H and for the special tube. These phase-space diagrams are like the Applegate diagrams of Figure 1 except that the slope representing the average electron velocity has been subtracted out. Notice that the bunch in the special tube is not as tight in the drift space preceding the output gap but contains more electrons when it passes through the output gap. Also, from the increased spacing between electrons that pass through the output gap 180° in phase away from the bunch center, one can see that the current density on the gap when the gap field is accelerating is substantially less, and fewer electrons take energy from the output cavity gap and waste it in the collector.

As a result of the higher intrinsic gain of five-cavity klystrons, it appears that one can achieve higher efficiencies in these tubes while still maintaining a gain that is adequate for use with modern solid-state drivers. The lower gain of four-cavity tubes makes such a gain-efficiency trade-off less practical.

In the future, the features of the special tube described above will be incorporated in our line of five-cavity 55 kW klystrons.

REFERENCES

1. E. L. Lien, "High-Efficiency Klystron Amplifiers", presented at the Conf. on Microwave and Optical Generation and Amplification, Amsterdam, The Netherlands, 1970.

2. R. S. Symons, "Klystrons for UHF Television", Proc. IEEE, vol. 70, No. 11, pp. 1304-1312, Nov. 1982.

3. R. S. Symons, A. Staprans, G. Caryotakis, "An Investigation of Optimum Bandwidth in Multicavity Klystrons", presented at the Tube Research Conf., Mexico City, D.F. Mexico, 1959 (unpublished).

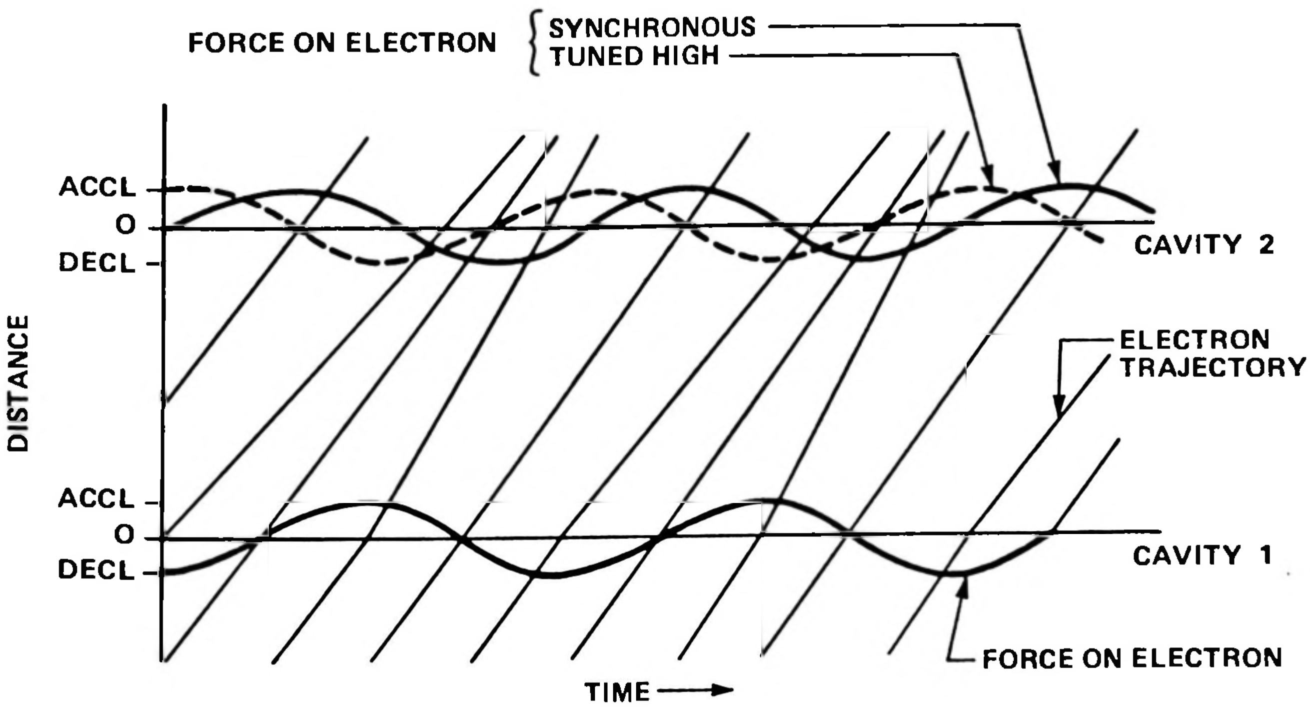


FIGURE 1. APPLEGATE DIAGRAM

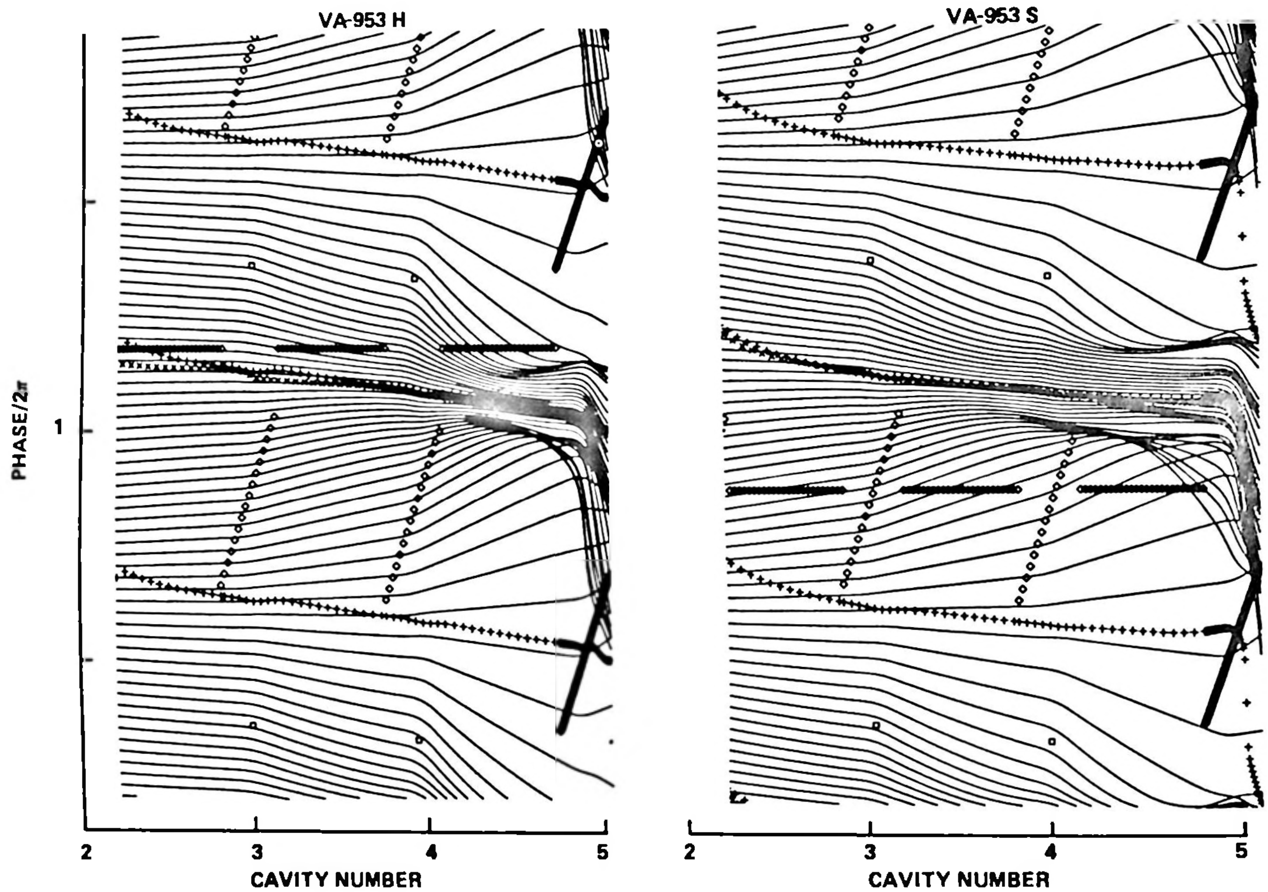


FIGURE 2. EFFECTS OF KLYSTRON TUNING ON BUNCHING

A DUAL CHANNEL SYMMETRICAL WAVEGUIDE SYSTEM

RICHARD E. FIORE & MARK A. AITKEN

COMARK COMMUNICATIONS, INC.

SOUTHWICK, MASSACHUSETTS

Introduction: The purpose of this paper is to describe in detail the design criteria utilized in determining the component design of a symmetrical waveguide system for the effective combining of dual television channels, the problems encountered in fabrication and test, and the final installed system operating characteristics.

The system under consideration is presently installed and operating on a tall tower structure in Cocoa, Florida, and utilizes square waveguide components. Two high power UHF television stations, operating on Channel 43 and 52, are effectively radiating through separate and isolated antennas through a single square waveguide transmission line feed system.

A complete comparison of the square waveguide operating characteristics relative to the conventional rectangular waveguide and circular waveguide is provided. Hopefully, sufficient information is presented such that the reader may fully understand why square waveguide was selected over circular waveguide as the transmission line medium. A treatment of the "Cross-pole" problems associated with polarized waveguide systems is presented and the methods utilized by various manufacturers to eliminate the unwanted side effects are discussed in some detail.

General Background On Waveguide Structures: The theoretical characteristics of rectangular, square, and circular waveguide have been well documented and utilized by Microwave System Design Engineers for many years. It is a well known fact that the most complete treatment of all types of waveguide theory and applications was completely documented in the well known and widely used "M.I.T. Rad. Lab. Series" of technical publications. This series of publications was the end result of a major effort extended during the pre-war years of

World War II. This dates the effort to the period of 1939 to 1946. To many of us, this represents a look back into history. To some of us it represents a time when we were first utilizing these publications as design standard reference books.

Possibly the best and most universally used handbook that was an outgrowth of the "M.I.T. Rad. Lab. Series" was a work compiled by Theodore Moreno, entitled "Microwave Transmission Design Data."[1] This writer can well recall one of his associates once saying that the only qualifications necessary to be a Microwave Design Engineer were to have a sharp file, a firm hammer and a copy of Moreno's Microwave Transmission Design Data Handbook. To be sure, if one were to be thoroughly acquainted with the contents of Moreno's handbook, one would certainly qualify as a practical microwave component design engineer. Mention of this reference text is brought to the readers attention since a selection of topics and time proven formulas relevant to this paper were obtained from it.

--- - ---

For purposes of this discussion, only symmetrical hollow pipe waveguide structures will be considered. Of specific interest is square and circular waveguide. It should be noted that all hollow pipe waveguide structures are subjected to the same physical laws in determining their individual basic operating parameters. It is only man who manipulates these physical qualities to suit his individual intent and purpose. For example:

(1) All uniform hollow pipe waveguide structures are high pass devices by nature. This means that they exhibit a minimum low frequency of operation at which all propagation ceases to exist. This frequency is designated as the cut-off frequency and is a function of the waveguide physical dimensions.

(2) All uniform hollow pipe waveguide structures are phase dispersive. This is to say that the wavelength in waveguide always exceeds the free-space wavelength and as the operating frequency approaches the waveguide cut-off frequency, the guide wavelength becomes increasingly greater and reaches infinity at low frequency cut-off.

(3) All uniform hollow pipe waveguide structures exhibit desirable impedance and insertion loss characteristics. Some guides offer exceptional characteristics depending upon physical dimensions for a given mechanical configuration and frequency of operation. Strangely enough, all hollow pipe waveguide structures exhibit decreasing insertion loss characteristics as a function of increasing frequency of operation; a characteristic that is exactly opposite to that of coaxial transmission line systems.

[1]Refer to list of reference articles in Appendix

(4) All uniform hollow pipe waveguide structures are capable of exhibiting undesirable operating characteristics. Some are able to be controlled by exercising proper theoretical conditions; some are inherent and difficult to eliminate on a practical basis and some will never absolutely be eliminated and the design engineer will have to live with the resulting consequences.

Consideration of items (1) through (4) are essential in developing the rational for selecting square waveguide over all other types for a dual signal propagating medium. It will be shown, hopefully to the readers satisfaction, that item (4) is the overriding consideration. A thorough understanding of item (4) by the cautious and knowledgeable consultant engineer will enable him to avoid the pitfall of selecting unique waveguide systems requiring sophisticated mode suppression techniques to provide long term, trouble free operation.

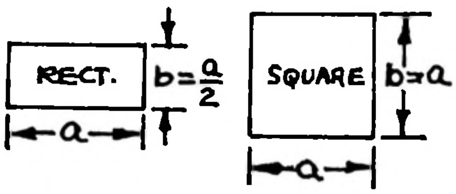
<u>Introduction To Basic Theory of Symmetrical Waveguide</u>: The requirement of signal isolation by geometric means and the subsequent propagation of orthogonal polarized signals immediately precludes the use of coaxial transmission line, rectangular, eliptical and single ridge waveguide structures. The necessary criteria for the required propagation requires axial circular symmetry about the transverse axis of the hollow pipe structure. Since actual pure single mode propagation is allowed by controlling physical dimensions, triangular cross sections are ruled out. This leaves square, circular, and a wide variety of the "agonal" family. Such devices having hexagonal or octagonal cross sections will work, but the utilization of even sided cross sections will result in the optimum circular cross section as a limit, hence, let us consider circular and square waveguide as the ideal structures for this discussion.

It is now essential to the discussion that the electrical and mechanical properties of circular and square waveguide be evaluated on a comparative basis.

<u>Operating Bandwidth</u>: Theoretically all waveguides can be operated from their lower limit cut-off frequency to their upper limit cut-off frequency, (fco), defined as that frequency, if exceeded, will allow the waveguide to support the next higher mode. Practically, for rectangular and square waveguides, it is advisable to limit the band of operation to:

$$\text{Low Frequency Limit} = 1.2 \times \frac{11{,}800}{2a}, \ (\text{MHz})$$

$$\text{High Frequency Limit} = 1.8 \times \frac{11{,}800}{2a}, \ (\text{MHz})$$



Where: a = Width of the waveguide in inches

b = Height of the waveguide in inches

For square waveguide $a = b$

Example: Compute the operating bandwidth of WR1500 waveguide

$$\text{Low Freq. Limit} = \frac{14,160}{30} = 472 \text{ MHz}$$

$$\text{Upper Freq. Limit} = \frac{21,240}{30} = 708 \text{ MHz}$$

Which yields an available useful bandwidth of 236 MHz.

--- - ---

Therefore, the usable bandwidth is computed to be 472 MHz to 708 MHz. If the interested reader checks further into bandwidths recommended by manufacturers for broadcast television purposes, he will discover that the usable bandwidth is further reduced to guard against:

(a) High attenuation characteristics at lower frequencies,
(b) Greater phase dispersion characteristics at lower frequencies,
(c) More rapid impedance change at lower frequencies,
(d) Possibility of moding and loss of efficiency at higher frequencies in complex waveguide systems.

It must be noted that the geometry of square waveguide is such that the dominant TE_{10} mode will be propagated in both the horizontal and vertical plane due to its symmetry,

--- - ---

Recently circular waveguide has been introduced to the broadcast industry since it exhibits very desirably low insertion loss characteristics and improved tower windloading factors. These characteristics compared to rectangular and square waveguide seems on the surface to be an answer to a "Maidens prayer". In order to obtain these desirable characteristics, one <u>must</u> utilize the waveguide in a region of its characteristic curves that allows more than one waveguide mode to propagate. This method is termed an "over-moded condition" and will be discussed in more detail later in this paper.

--- - ---

Admittedly there are various methods of mode suppression available to the design engineer, however the merit of these techniques should be critically examined. The one electrical design consideration that severely limits the practicality of symmetrical waveguide systems for broadcast purposes is the ability of circular or square waveguides to cross couple to their dominant modes and set up fields that are orthogonal to the field of the dominant mode. All of the literature available seems to indicate that there essentially is no complete solution to this problem except to provide a mechanism for extracting and dissipating the energy. This method of course requires that the cross-pole component be accounted for in the published insertion loss data. The ITT Reference Data For Radio Engineers[1] states that, "In practice waveguides capable of supporting propagation of dual-polarized

[1]Refer to list of reference articles in Appendix

signals are subject to cross-pole problems. In practice, wall losses, surface irregularities and unequal transverse interior dimensions give rise to cross-conversion of the dominant propagating mode of the waveguide. Certain design considerations are given to minimize this effect. These considerations deal with the frequency band existing between cut-off and the first higher order mode frequency of propagation. Specifically, it is recommended that the lowest frequency of operation be 25% above the dominant mode cut-off and well below the next higher order mode frequency." The definition of "Well below" is left to the reader but would suggest 25% below as a minimum. Further clarification of the subject is given by Moreno[1] on Page 118 of his handbook. This reference clearly states that, "In general, if a circular waveguide is deformed, the wave being transmitted will split into two components that proceed down the pipe with different phase velocities and different attenuation. This instability will be found in all cases except the following:

1. When deformation is along an axis of symmetry of the wave.
2. When the wave has circular symmetry (e.g., TE_{01} or TM_{01} wave).

Under these conditions, it can be expected that even the most precise waveguide structures available will be subjected at some time or place to environmental conditions that will cause undesirable system problems."

In the case of circular waveguides, the waveguide useful bandwidth limitations are more critical. The reason for this is that for a given diameter of circular waveguide the total bandwidth existing between cut-off and the first higher mode limiting frequency is only a fraction of the bandwidth available to square waveguide users. The available bandwidth is defined by:

$$f_{LCO} \text{ (Lowest Freq.)} = \frac{6915}{D''} \text{ (MHz)}$$

$$\text{(Highest Freq.) } f_{HCO} = \frac{9039}{D''} \text{ (MHz)}$$

Using the same criteria as used with square waveguide concerning the 20% frequency buffer zone above the low cut-off frequency and below the next higher moding frequency, there would be no usable bandwidth left. Thus, it is up to the design engineer to determine what technical liabilities he will be subject to and make his decision accordingly. This writers experience has shown a good rule of thumb to be to limit the usable bandwidth of any customized circular waveguide to:

$$\text{Lowest Frequency} = \frac{7500}{D''} \text{ (MHz)}$$

$$\text{Highest Frequency} = \frac{8500}{D''} \text{ (MHz)}$$

Example: Determine the optimum size circular waveguide for Channel 28 (557 ± 3 MHz)

[1]Refer to list of reference articles in Appendix

(a) Determine diameter D in inches from

$$D'' = \frac{8200}{f_{CB}} : \quad f_{CB} = 557 \text{ MHz}$$

$$D'' = \frac{8200}{557} = 14.72'' \text{ diameter circular waveguide}$$

Which size according to available published data, indicates an attenuation figure of approximately 0.065 dB/100'.

This data proves to be no better than that data published for WR1600 rectangular waveguide. It should be pointed out that the published data for circular waveguide encompasses frequencies of operation above the frequency of the first higher order mode and up to the frequency of the second higher order mode. This places the user of this data in a very precarious position unless he is fully cognizant of the pitfalls and problems associated with higher order modes in circular waveguide.

Circular waveguide as compared to square waveguide will propagate the dominant TE_{11} mode in an infinite number of planes along the transverse axis instead of being confined to the vertical and horizontal planes of a properly designed square waveguide system supporting the dominant TE_{10} mode.

Insertion Loss Characteristics

Typically, any comparison of insertion loss comparison calculations can be made to furnish any desired result based upon the physical assumptions made. For this reason it is essential that certain basic assumptions be made that allow the comparison of "Apples to Apples" as applied to waveguides.

Contained in the appendix of this document is an analysis of the insertion loss of circular waveguide as compared to square waveguide; under the assumptions that:

(1) The same design frequency is used.
(2) Basic waveguide dimensions allow only the dominant mode to propagate. This condition is imposed purely for purposes of determining the most logical selection of the type of waveguide to be used in a dual mode waveguide system such as that under consideration in this paper.
(3) Pure copper conductors are utilized throughout.

The calculations clearly indicate under the conditions imposed that square waveguide is at least 25% more efficient than circular waveguide of the same size. This statement may cause some eyebrows to be raised; however, under the assumed conditions, it stands as a valid and a defensible statement.

Power Handling, Impedance and VSWR Characteristics Power Handling: Unobstructed waveguide be it rectangular, square or circular will handle extremely high power levels at the designed frequency of operation. The ability of waveguide to handle power only becomes limited when such items as resonant structures, matching (VSWR) devices and screw tuners (threaded rods) are scattered about with wreckless abandon. One capacitive tuning slug placed in the maximum electric field position of a waveguide can reduce the voltage flash over by a factor of 10:1. Even under these conditions, waveguides will more than handle any broadcast television maximum power levels with a substantial margin of safety.

Impedance: Waveguide impedance is the underlying reason why waveguide is more efficient than other transmission line types and account for its high power handling capability. The universal expression for waveguide impedance is given by:

$$Zog = 240\pi \frac{b}{a} \frac{\lambda g}{\lambda o} \quad \text{(rectangular waveguide)} \quad b = 0.5\, a$$

Rectangular waveguide impedance decreases as frequency increases exhibiting values of 700 ohms to 450 ohms over its useful range. Thus the principal of transferring large amounts of energy long distances follows the high voltage low current philosophy utilized by commercial utilities. In this case, large amounts of radio frequency energy propagate through high impedance devices with considerably less attenuation. For square and circular waveguides, the characteristic impedance is normally twice that of standard rectangular waveguide. Due to the much larger characteristic impedance, (Zo), which is determined by physical parameters, small impedance fluctuations due to bends, ripples or oil canning as a result of the fabrication process has much less effect resulting in a smoothing of the VSWR characteristic.

System VSWR: For some unknown reason, a cursory examination of most waveguide manufacturers sales literature reveals the interesting fact that extensive tuning of rectangular waveguide systems utilizing individual external waveguide tuning mechanisms are required to achieve optimum VSWR results on tall tower structures. This writer finds the statement hard to believe, since every system designed and installed according to the theory about to be discussed has resulted in excellent operating results with no external tuning mechanisms of any kind applied to the vertical run.

There are three essential design considerations that must be understood and precisely adhered to in order to develop waveguide systems for tall tower structures, they are:

(1) Strict attention to maintaining minimum physical dimensional tolerances during the actual fabrication process is essential,
(2) Techniques to eliminate unwanted axial twist build-up in waveguide structures must be developed,
(3) Each length of the vertical run must be precisely machined to a fixed dimension; an odd multiple of waveguide quarter wavelengths.

Design considerations for selecting the optimum system configuration for the simultaneous transmission of two television channels efficiently through a single tall tower waveguide system.

Up to this point in the discussion, basic theoretical concepts have been presented in order to acquaint the reader with the characteristics of symmetrical waveguide. Now with the preliminaries out of the way, the moment of truth is at hand as to what type of waveguide structure would perform the required task most effectively. The basic system requirements to be met are as follows:

- Select a waveguide structure that will allow orthogonal mode propagation with maximum isolation.
- Select an efficient, low loss, structure.
- Select a waveguide structure with sufficient bandwidth that will allow optimum VSWR characteristics for two channels separated by a minimum of sixty (60 MHz).
- Select a waveguide structure with mechanical integrity and as low a windload factor as possible.

Required Characteristic	Circular Waveguide	Square Waveguide
Insertion Loss	Good To Excellent	Good To Excellent
Theoretical Orthogonal Mode Purity	Poor To Good With Mode Suppression	Excellent
Mode Free Bandwidth	Poor	Excellent
Structural Integrity & Windload	Excellent	Good

The above chart was determined from the previous discussed component system considerations. As can be expected, for a "First of its type" system, the degree of success in achieving an operational dual channel system would favor the use of square waveguide. Therefore the balance of this paper will be concerned with the design criteria basic component design and final system operational characteristics of a square waveguide system.

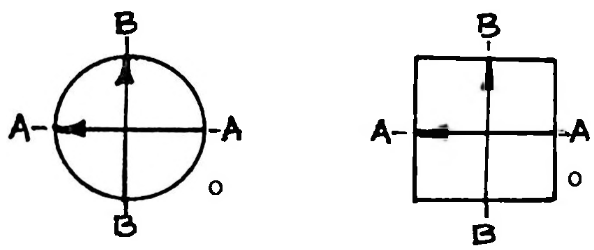
System Design Objective: To design and develop a single run of transmission line to propagate two separate and distinct high power UHF broadcast television channels up a tall tower structure without the utilization of frequency sensitive channel combining equipment at the input and output terminals requires the transmission line to:

(a) Accept polarized signals and maintain the plane of polarity.

(b) Maintain the isolation of the polarized signals relative to each other such that moding and cross-pole coupling does not cause degradation of picture quality and system efficiency.

Polarized non-coherent frequencies can only be propagated through hollow wall waveguide that is, as a minimum, symmetrical in two orthogonal planes normal to the axis of propagation. The two most desirable types of waveguide that will perform the required task are circular and square waveguide supporting dominant modes in two planes as shown in fig. 1. As stated previously, theoretically both types of waveguide should exhibit infinite isolation between the two modes. In real life such a condition is impractical and as the signals propagate down the length of the transmission line, mechanical forces take over and degrade the level of isolation as a function of total length. This condition in addition to flange interface reflections and cross-pole due to higher order moding problems are all summed up to form a single resultant amplitude and phase distortion problem. The amplitude problem takes the form of increased insertion loss; the phase distortion problem results in a critical group delay problem relative to allowable group delay of the transmitted signal. The cause and effect of these problems are well documented and the reader is encouraged to obtain copies of the listed referenced articles (4, 5, 6) listed in the appendix for more information on the subject.

fig. 1



TE_{11} (AA, BB) TE_{01} (AA, BB)

°Direction of Propagation into paper.

Once again so that the reader will fully understand the logic behind selecting square waveguide for the propagating media of the system under consideration, the reasons are clearly stated.

(1) Both circular and square waveguide are susceptible to cross-pole generated signals. The fact that the square waveguide fixes the plane of orthogonality for the two signals as compared to circular guide where the planes of polarization are free to rotate relative to one another is of considerable importance.

(2) The mode free percentage bandwidth for square waveguide is much greater than circular waveguide, thereby decreasing the probability of moding problems for two channels separated by a significant percentage of the available waveguide bandwidth.

(3) Since propagation in the square waveguide is limited to the TE_{10} mode by design and the major planes of propagation are defined, it is predicted that control of the magnitude of the cross-pole component will be more easily accomplished. In the case of circular waveguide, a cross-pole component can exist in any plane depending upon the rotation of the plane as function of forces acting along the path of propagation. In a dual signal system both signals start out orthogonal, however as they travel along the waveguide both signals will rotate to some degree in a random fashion, absolute control of the cross-pole components is then a difficult problem to overcome.

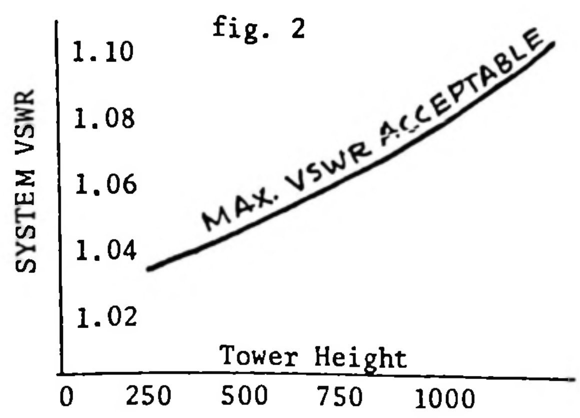
(4) The insertion loss of square versus circular waveguide is essentially the same and not a consideration.

(5) The mechanical problems in fabricating and in supporting square waveguide on a tall tower structure are essentially the same as those of circular waveguide. Admittedly square waveguide would be subject to slight twist as compared to circular; however, these problems have been expertly examined by leading structural consultants and suitable techniques on restraining motion have been designed and are available for use.

Design criteria for optimizing square waveguide dimensions to propagate two separate TV channels efficiently.

As previously discussed in this paper all single channel bandwidth waveguide systems are usually designed to efficiently operate about the center frequency of the television channel bandwidth under consideration e.g. fo ± 3 MHz. The usual procedure to derive an optimum design is to first select the largest size waveguide that will meet all propagation requirements including higher order moding considerations; second, determine the optimum per-unit piece length to minimize flange interface reflections and produce the lowest realizable bandwidth VSWR characteristic. It has been this writers experience that if this procedure is rigorously followed, waveguide systems manufactured to EIA tolerance specifications for various vertical heights as shown in fig. 2 will exhibit excellent system VSWR characteristics without the requirement of external tuning devices. In the case of a waveguide designed to operate efficiently on two different channels the optimizing procedure is somewhat more complicated, but certainly mathematically realizable.

fig. 2



Design criteria for launching two independent television channel signals into a square (or circular waveguide) system.

As is generally the case in any system design, there is at least one key component that allows the system to be realizable. In the case of a dual mode system such as that under discussion, this component is the orthogonal mode transducer.

The orthogonal mode transducer in simple terms is an impedance transforming device which allows the low impedance rectangular waveguide to be coupled efficiently to the high impedance square waveguide yet maintain signal isolation between channels. Since the signals are to be combined at the base of the tower, they in turn must be separated at the top of the tower when channelized antennas are employed. Since this is the case under consideration, the system as designed requires two orthogonal mode transducers one at the input and one at the output to be a practical system.

The basic design criteria of orthogonal mode transducers has been common knowledge in the field of microwave component design for many years. This writer participated in a lengthly and complicated design of broadband orthogonal mode transducers utilizing circular waveguide while toiling in the ivory halls of the Radio Corporation of America some odd thirty years ago. This work was completely documented in a published technical paper[1] at the culmination of the project. The sum and substance of the design project was that broadband orthogonal mode transducers could be developed which exhibited exceptional broadband frequency isolation and low VSWR characteristics. Much of the information derived from this design effort was later utilized in developing a product line of very high power variable waveguide attenuators and specialized polarized antenna feed systems.

Having this prior experience to generate the required technical know-how it was a relatively simple task to generate a conceptual design of the orthogonal mode transducer required for the symmetrical waveguide system under discussion.

As is the case in all conceptual design projects, the design engineer is usually confronted with several unpredictable design flaws that if not solved will quickly terminate the program. Such was the case experienced in this instance. Two problems developed that at first seemed insurmountable. The two problems stated in simple terms were:

1 Refer to list of reference articles in Appendix

1. Internal structures of the orthogonal mode transducer caused unwanted resonance characteristics to be evident in the operating waveguide bandwidth.

2. Mode conversion characteristics of the waveguide structures utilized caused periodic pertubations in the Amplitude and Phase transfer functions of the passive system components under test.

The first of the two problems, the resonance phenomena was determined to be due to critical septum lengths that were essential to the operation of the orthogonal mode transducer. It is common knowledge to those technical inclined that the presence of such resonances guarantee undesirably high dissipation losses and unacceptable amplitude and phase transfer characteristics. The proper solution to solving this problem was found to be available through optimizing the dimensions of the biforcating septum of the square waveguide structure (See fig. 3). It was first discovered that the unwanted resonances could be moved outside of the critical channel bandwidths under consideration. Later on it was discovered that by extending the septum through the transforming devices that the resonances could be eliminated altogether. fig. 3 shows the basic geometry of the orthogonal mode transducer with the critical septum dimensions indicated for the readers edification.

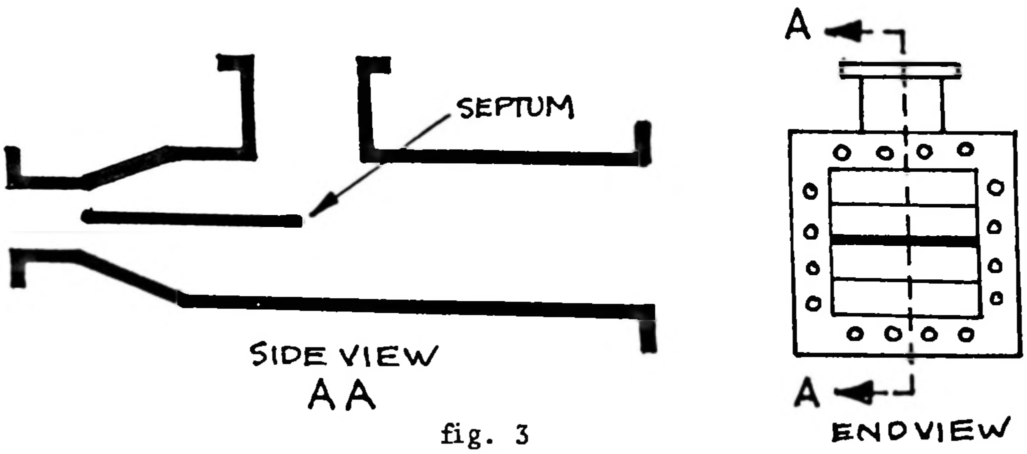


fig. 3

The second of the two problems, the mode conversion characteristics of symmetrical waveguide structures be they square or circular waveguide could easily develop into a subject for a doctoral thesis for some aspiring graduate student. Since this writer does not have these aspirations, a more profound attempt will be made to document the cause, effect and remedy of the mode conversion problem.

A considerable amount of information has been written concerning the utilization of over-moded waveguide structures to obtain very low loss waveguide system characteristics.[1] In all of these articles, the primary concern is highlighted to be the elimination of the mode conversion process through a number of available techniques, yet realize that the best of available techniques are not loss free and those losses must be accounted for when the ultimate system analysis is made.

The factors that determine the absolute amount of energy loss due to the mode conversion process of a symmetrical transmission line system are:

1. The basic physical configuration of the symmetrical waveguide structure which determine the over-moding characteristics of the system.

2. The technology utilized in designing the optimum mode transducer for the system under consideration.

3. The precision required to fabricate the final system components such that the completed components are as perfectly symmetrical as required by theory.

4. Waveguide component tolerances as prescribed by the method of manufacture.

5. Flange interface problems due to mechanical dimensions and fabrication techniques.

6. The total length of the transmission line system.

7. Methods utilized to suppress undesirable mode characteristics.

8. Effects of system environment upon long term stability of final system.

For purposes intended, let us consider the factors listed.

The first factor, Physical configuration, determines the ability of any symmetrical waveguide structure to propagate the dominant mode only or is purposely designed large enough to allow propagation of one or more higher order modes. In the case at hand, this would be:

	Dominant Mode	1st Higher Order Mode	2nd Higher Order Mode
Square	$TE_{10}^{\square}$	$TM_{11}^{\square}$	$TE_{11}^{\square}$
Circular	TE_{11}°	TM_{01}°	TE_{21}°

Currently, all of the most recent published technical articles and manufacturers data concerning circular waveguide, stress the use of circular waveguide utilized in the over-moded state. This means

[1] See Appendix for technical listing of references.

that in order to achieve the desired low system insertion loss (dB attenuation per unit length), the circular waveguide must by necessity be physically designed to support one or more higher order modes in addition to the dominant TE_{11}. Once the design engineer commits to this criteria, the system design becomes much more complex.

To appreciate the complexity of the problem, the interested reader need only read the referenced applicable articles listed in the appendix of this paper. To outline some of the more critical problems, the need for:

- Ultra precision fabrication techniques are necessary.
- Utilization of mode suppressors absolutely required.
- All losses due to mode conversion and mode suppression must be accounted for in the overall system.
- For systems where mode conversion at both ends of an over-moded waveguide single-mode or rectangular waveguide or eliptical waveguides, a mode suppressor must be built into the transmission line system to avoid the peak losses due to trapping of higher order modes at the transition at each end of the transmission line system.
- Long term environmental conditions should be examined for any frequency sensitive mode suppressing techniques utilized.
- The mechanical strength of the over-moded waveguide components should be such that suppressed trapped resonant spikes do not reappear when the transmission line is subjected to all forms of external forces and environmental conditions.

Fortunately for the telecommunications industry, the systems are short in physical dimensions (less than 200 feet) and the components lend themself to precision electroforming techniques and heavy wall tubing materials for structural ridgidity. Additionally, the mode suppressing techniques need only absorb at best very low power levels. When one thoroughly familiarizes himself with the problems that must be resolved with small diameter tubing (4" dia. or less), one must recognize the magnitude of the problem when the techniques are applied to circular waveguide systems utilizing waveguide diameters of from 12 to 20 inches.

With regard to precision components the electroform process is not applicable to large waveguide structures. Standard large waveguide fabrication techniques are limited to certain mechanical tolerances. The mechanism for rolling, seam welding and flange welding are fraught with distortion and stress relieving problems that cannot be overlooked. Claims of super precision devices should be discounted to what is possible to achieve and maintain as a function of time and external forces encountered.

The utilization of absorbtive vane type broadband mode suppressor devices are presently not applicable as high power broadcast television waveguide components. This statement does not preclude that high power-absorbtive mode suppressors will not be available in the near future.

The utilization of external waveguide deforming mechanisms to set up equal and opposite mode suppressing signals only serve to increase overall attenuation characteristics due to mode losses. This technique cannot be viewed as an optimum long term solution to eliminate moding problems for tall tower waveguide systems designed to propagate one television signal, let alone two signals under severe environmental conditions.

It is difficult to comprehend how even the most precise large diameter (12" to 28") circular waveguide will not be subjected to some mechanical deformation of an elyptical nature (a problem which most all of the technical information available sites as the basic problem with circular waveguide systems. When one considers that the intended system must operate from -25°F to +115°F with wind velocities of up 100 MPH when mounted on tall tower structures (800 Ft to 1600 Ft), which by themselves subject the waveguide system to severe stresses as a function of their physical construction, one must conclude that extreme caution should be utilized in selecting the type of waveguide system to be employed.

When one considers the arguments presented and substantiated by documented data, even the most optimistic of microwave system design engineers would be reluctant to select circular waveguide operating in the over-moded method for the purposes intended in this paper. Thus the only available alternative left after considering the problems associated with circular waveguide was to select square waveguide as previously concluded for the transmission line system. The reader is cautioned that square waveguide can also be operated in the over-moded condition, however, if the design criteria previously established is utilized, that is limiting the waveguide size to support <u>only</u> the dominant mode, the designer will only have to contend with mode conversion from the $TE^{\square}_{10}$ to $TE^{\square}_{01}$ which reduces to a dominant mode cross-pole problem without the complexities of mode rotation and loss of efficiency due to mode suppression and signal extraction techniques.

One other factor that cannot be passed over lightly is what this writer terms as the "futures" of waveguide television systems. If for example the mode suppression techniques available are found to be limited and undesirable signal distortion characteristics are reinserted into the swept bandwidth response of the television transmitter signal, what profound effect will it have on proposed future improvements in broadcast television systems. For example, what would be the consequences of periodic envelope delay and amplitude spikes upon the proposed multiple channel sound and high definition (1000 line) video systems. This leaves food for thought and encourages restraint on the part of systems designers to offer a product that will meet the needs of future broadcast television transmitter systems.

Component Integration and Transmission Line System Test:

Upon completion of the orthogonal mode transducer design and making the decision to utilize single mode square waveguide as the transmission line mechanism, it was a relatively simple task to reduce the prototype to functioning components. The completed production model orthogonal mode transducer exhibited a maximum VSWR of 1.03 to 1 as viewed at the input orthogonal ports with the square waveguide output port terminated in a matched load a VSWR of 1.02 to 1. The back to back electrical test of two orthogonal mode transducers (See fig. 4) resulted in tunable VSWR maximums of 1.03:1 with orthogonal signal isolation of greater than 34 dB for each channel bandwidth of interest. The assembly was also checked for any sign of moding problems or undesirable resonances with no evidence of any undesirable signal distortions.

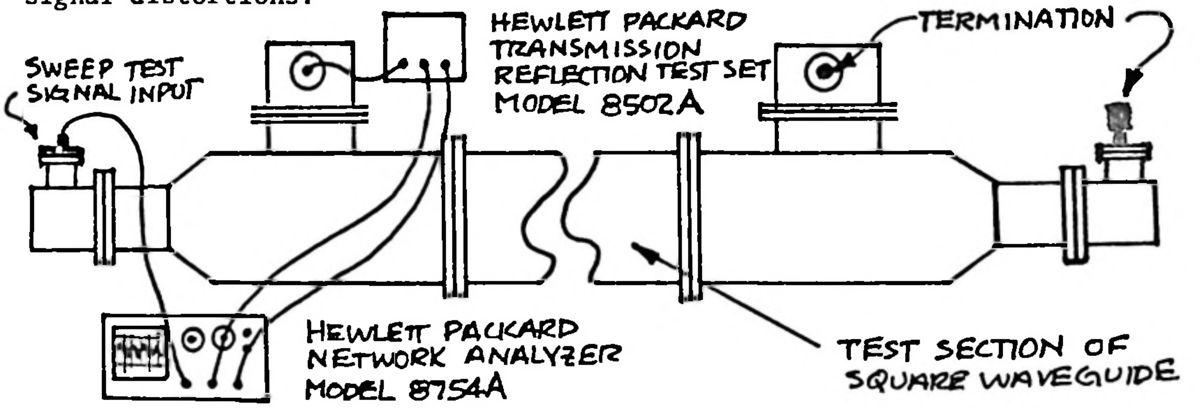


fig. 4

The next phase of the test procedure was to start inserting the production run of fabricated square waveguide sections between the orthogonal mode transducers and recording the electrical parameters as a function of total system length.

Prior to commencing this process, a mechanical (random sample) inspection of ten sections of production fabricated square waveguide (11.6" x 11.6" x 144") to check for manufacturing accuracy. The inside dimensions were checked at two foot intervals along the Z axis of the waveguide (12 foot dimension). A plot of deviation from specified dimension is shown in fig. 5 for the random samples selected.

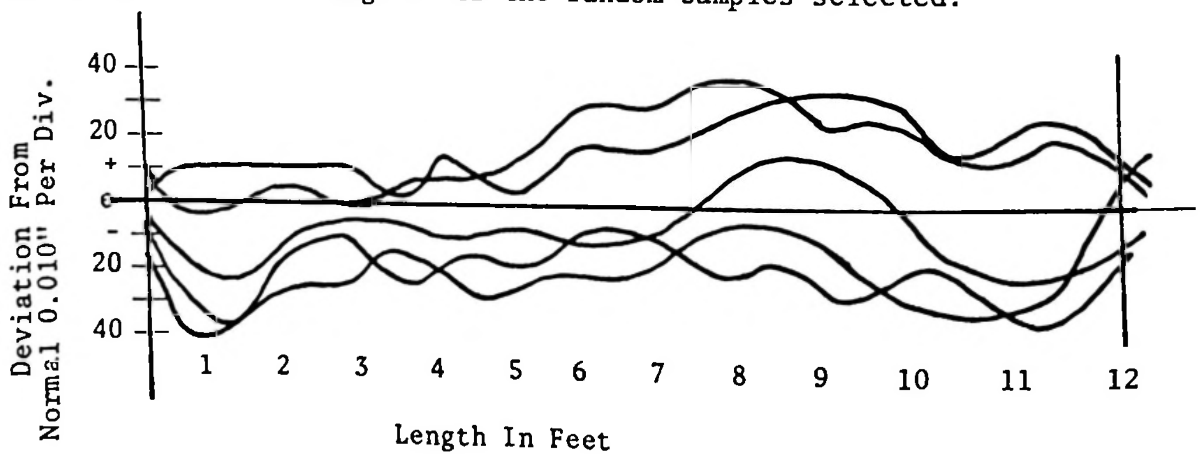


fig 5.

It is quite evident from the sample data that square waveguide can be easily manufactured to the standard EIA specified dimensions associated with the WR series rectangular waveguides. Additionally, the square waveguide was checked for "squareness" or possibly better termed the parallelogram effect. Again the tolerances were found to be excellent for a first time through production run of waveguide components.

Having ascertained the precise physical nature of the fabricated square waveguide, the orthogonal mode transducers were positioned on a flat factory floor surface and the square waveguide pieces were stacked one at a time between the transducers and the electrical parameters checked. Almost immediately it was evident that the isolation characteristics between the two orthogonal mode transducers became progressively worse as the number of pieces was increased. For those readers familiar with orthogonal mode transducers, the isolation characteristic is a $COS^2\theta$ function on a power basis as Shown in fig. 6. Thus for each section added, any cross-pole generated in each section should show an accumulative effect and cause the isolation to deteriorate from some level in excess of -36 dB. It was very evident that the degradation was broadband, since channel to channel isolation was decreasing approximately the same magnitude at the two channels of interest. It was interesting to note that system VSWR as a function of length remained quite acceptable and was relatively independent of length. As a matter of fact it looked quite smooth and lacked the typical CUSP type impedance display when viewed using the Hewlett Packard 8754A Network Analyzer in conjunction with the Hewlett Packard transmission reflection test set Model 8502A. It is important to point out that impedance and transfer function measuring equipment of this type is absolutely essential to measuring these parameters. This is due to the fact that any swept frequency measuring device must be cautiously examined such that the gain bandwidth product of the system is not violated and sharp discontinuity peaks are not truly displayed due to their rapid frequency versus amplitude excursions.

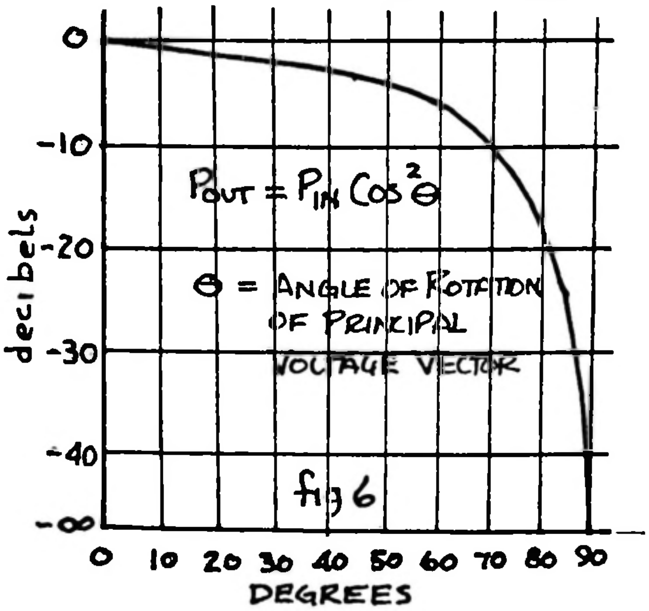
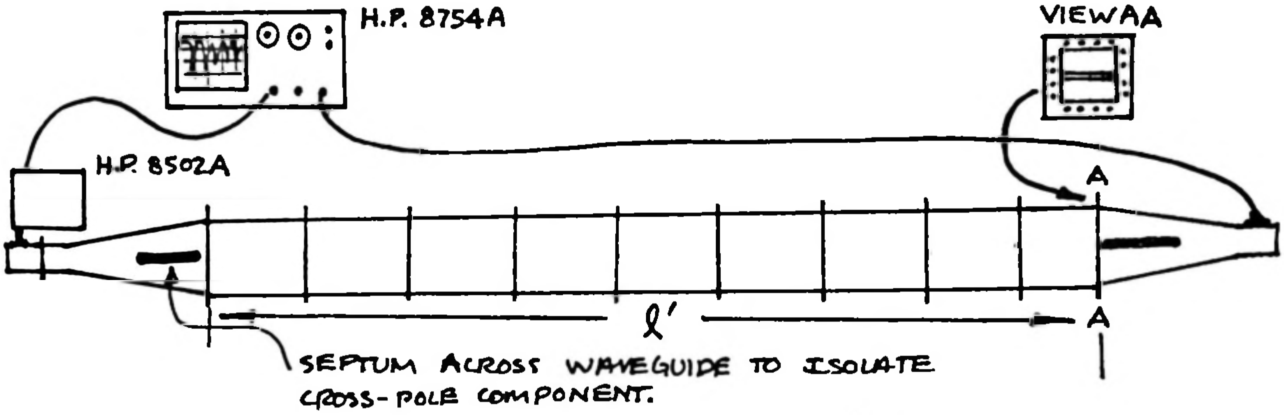


fig 6

Upon experiencing the decreasing isolation phenomena due to cross-pole conversion, it was decided to stop the process and investigate the problem to determine a satisfactory solution. Having anticipated the possibility of such a problem before hand, the orthogonal mode transducers were replaced with two rectangular to square waveguide tapered transitions built specifically to investigate such effects. The square waveguide was biforcated in one plane so that cross coupled components were electrically shorted depending upon the plane of polarization being investigated. This was done to insure that only cross-pole magnitudes free of second order effects would be electrically visable for evaluation.



$$\Delta f = \frac{492}{\text{DISTANCE } \ell \text{ IN feet}} \times \frac{\lambda_g}{\lambda_o}$$

fig. 7

fig. 7 represents the evaluation technique used to observe the cross-pole phenomena. fig. 8 indicates the spiking distortion in both phase and amplitude transfer function measured characteristics.

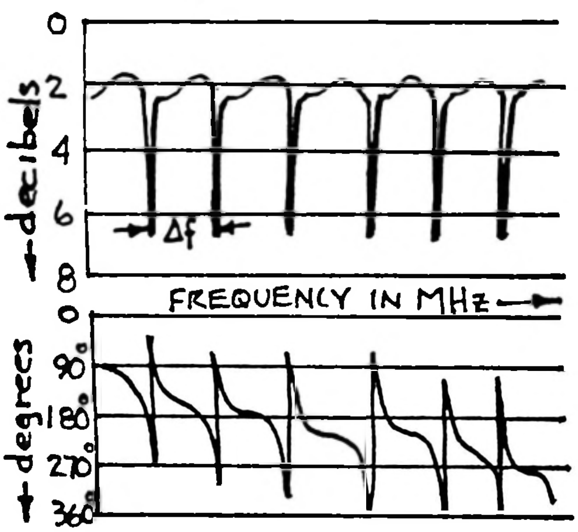


fig. 8

The most simple technique for rectifying the cross-pole spiking problem was to resort to some type of deformation process such as that thoroughly described by the Andrew Corp. in published literature for their over-moded circular waveguide systems designed for telecommunications purposes.[1] It was quickly determined that by simply exerting a mechanical force component across the diagonal dimension of the square waveguide, the cross-pole spiking problem was practically eliminated. It was interesting to note that the entire bandwidth of interest (640 MHz to 710 MHz) was compensated with equal effect. This situation seems to differ drastically from the effect noticed when the same technique was utilized with over-moded circular waveguide systems. In such cases a narrow band distortion free frequency window is created. Additionally, some random spiking of rather limited amplitude was noted in the actual passband of interest. (See fig. 9). This situation would leave one to believe that the magnitude of the spiking problem was more complicated due to the circular symmetry, thus allowing many more mode components to generate, each contributing to the overall total problem.

[1]Refer to material listed in index of referenced technical articles

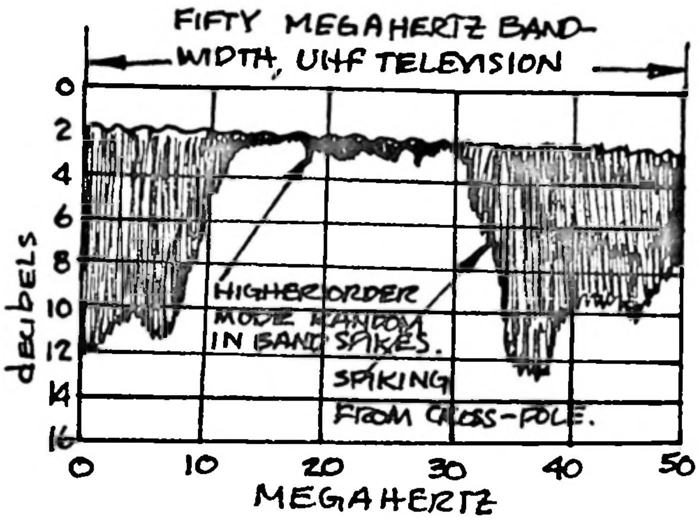


fig. 9

This window effect was not noticeable in the response characteristic of the square waveguide system when fed with the orthogonal mode transducers with cross-pole compensation or without. From this fact, it was concluded that all symmetrical waveguide systems should be excited with orthogonal mode transducer, compensated for cross-pole, and factor the additional losses into the total system insertion loss figure for F.C.C. license application purposes.

The Final System Configuration: With the completion of the back to back orthogonal mode transducers and the square waveguide stacked for a linear distance of five hundred feet, one would feel confident with overall system performance data as shown in fig. 10.

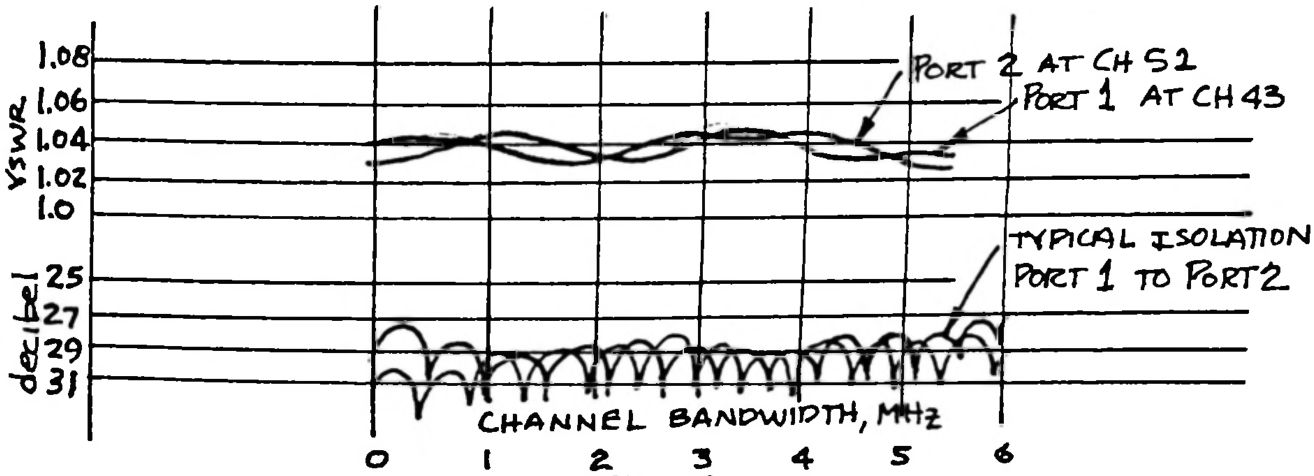


fig. 10

Indeed there is a certain degree of satisfaction in obtaining good performance data concerning the overall system, however, as is always the case, there is almost always a "slip between the cup and the lip" in most complex systems designs. In this instance there was one problem that required a proper solution.

As the data indicates in fig. 10, the isolation characteristic between operating channels can be identified to be a nominal 28 dB when the system is properly adjusted. It was quickly determined that the isolation varied ±3dB about the nominal when the system was subjected to extreme environmental conditions. For example the waveguide structure sitting in the hot sun can accumulate a surface temperature of upwards of 115°F, suddenly a cold front thunder shower moves through the area reducing the waveguide surface temperature to 50°F due to upper atmosphere physical conditions. If such temperature extremes do influence the system electrical properties, does not it make sense to compensate the system for long term variations and provide additional filtering to remedy any problems caused by periodic isolation variations.

To be perfectly clear on the subject, tests were conducted with dual system isolation as low as 23 dB. No detrimental system effects were noted on individual transmitter performance specifications. This is reasonable since the two channel frequencies are some fifty megahertz (50 MHz) separated and the selectivity of the klystron cavities provides sufficient roll-off isolation to prohibit any non-linearities. Unfortunately, the fault sensing systems are not so frequency selective and record such signals as high VSWR values on transmitter reflectometers. Since thirty decibels of isolation calculates to be an indicated VSWR of 1.06:1, a reading that would completely mask the actual transmitter system VSWR monitoring circuit protection capabilities; it became evident that some method of filtering to improve system isolation to forty decibels (40 dB) under the most severe environmental conditions was an absolute necessity. Additionally, any filtering utilized had to be effective over a wide temperature extreme. The solution to the problem was relatively simple to solve when properly analyzed. By building dual notch filters for both the visual and aural carrier frequencies of both channels and stagger tuning for maximum bandwidth about carrier (See fig. 11), the carrier frequencies will always be attenuated a minimum of 15 dB over the lowest isolation figure. This computes to be -40 dB and reduces the deflection of the VSWR meter circuits due to isolation to a negligible level. The final filter assemblies were incorporated into the transmission line feeding the orthogonal mode transducer at the bottom of the tower. The reason for positioning them at this point was to eliminate the possibility of any ghost signals due to phase differences of reflected signals. The final system configuration is shown in Drawing A500050 on Page 27 of this report.

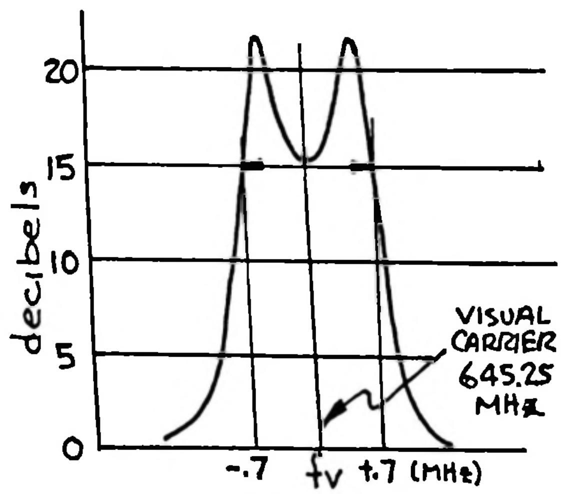
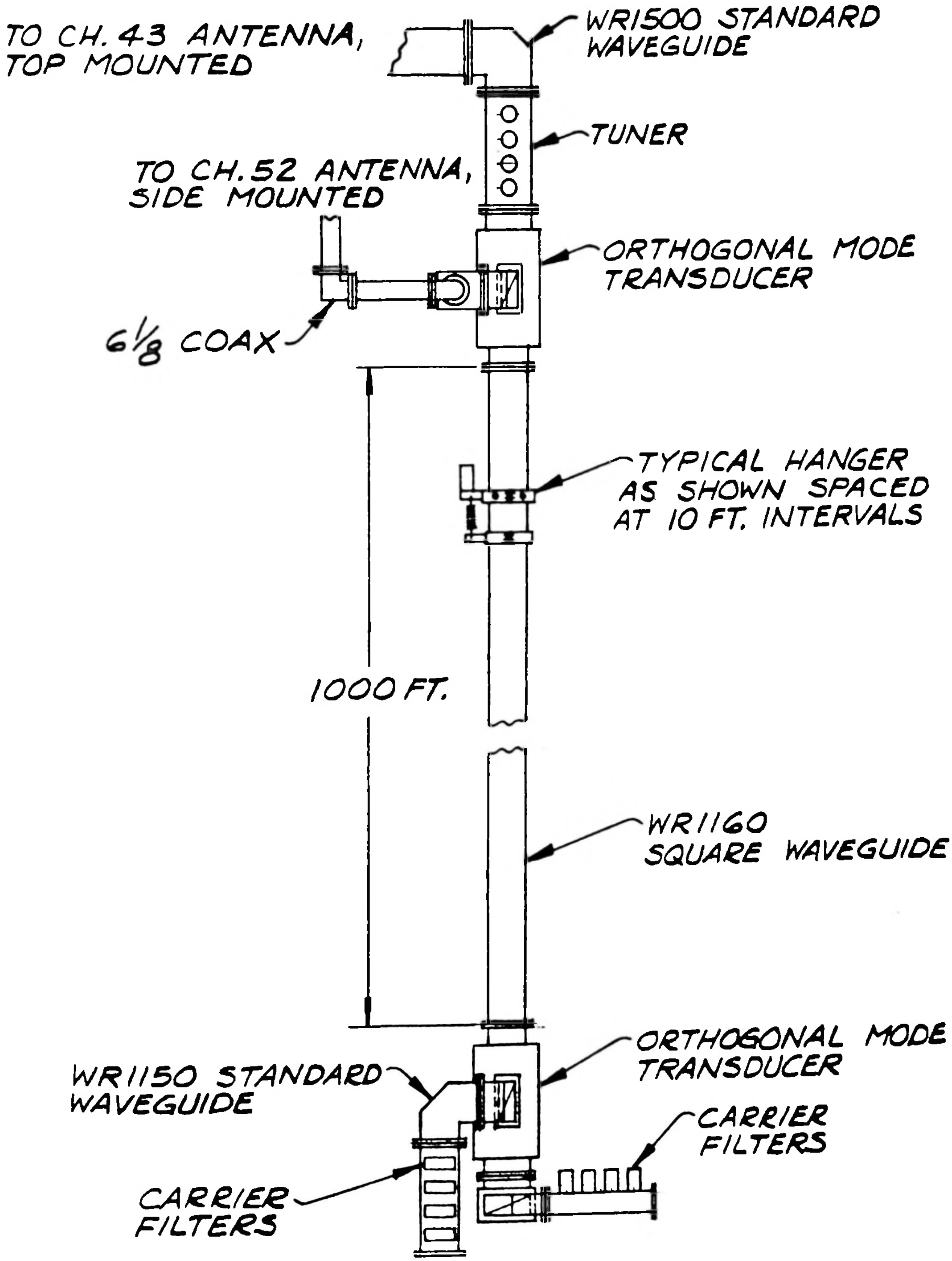


fig. 11

Conclusions: The fundamental conclusion that one can draw from the system design described by this paper is that it is entirely possible to utilize a symmetrical waveguide system to feed two channelized antennas with excellent electrical characteristics. The question that an attempt has been made to answer, is it more realistic to utilize an under-moded square waveguide with its slightly larger windloading factor as compared to circular waveguide or to use over-moded circular waveguide and live with the possibility of unwanted side effects associated with the over-moded principal of operation. Certainly when one takes into consideration that essentially little if any improvement of insertion loss is achieved in a properly terminated over-moded symmetrical waveguide system due to its inherent complexities the question becomes moot. Operation of the square waveguide in what might be called the "under-moded" condition as relative to the "over-moded" condition does in truth restrict the magnitude and phase orientation of the principal traveling wave "E" field configuration. In the case of circular waveguide, where the cross-pole and over-moded components can



A500050 : SQUARE WAVEGUIDE SYSTEM

wander about the axis and propagate at various phase velocities, it seems difficult to comprehend how optimum values of fundamental signals can be coupled out to the antenna structures through any kind of a coupling device. Again, in the case of square waveguide, orthogonality of dual signals is maintained, thus isolation between the two antennas is maintained. It seems that in circular waveguide, the final orientation of the fundamental "E" field configurations could be easily something other than orthogonal thereby causing less than optimum power being transferred to the proper radiating structures. Until the mechanism for maintaining orthogonality and isolation are more thoroughly investigated in an actual dual channel circular waveguide system, the logical approach to dual signal systems favors the less problematic square waveguide configuration.

In the evaluation period prior to completing the final system design, it was determined from actual field and laboratory observations that no sensational or phenominal improvement in total system insertion loss was going to be in evidence by utilizing circular or square waveguide in any mode of operation. By the time component loss factors due to dominant mode cross-pole, higher order mode conversion, isolation factors and component interface losses are collectively accounted for, the absolute value of system insertion loss was found to be only a small percentage improvement over optimized conventional rectangular waveguide systems. Naturally it is essential to remind the reader that drastic improvement of insertion loss was not the ultimate goal of this exercise. The propagation of two signals in one transmission line system could not be achieved in a conventional transmission line system without the use of complex channel combiners and broadband antenna structures.

To terminate this dissertation with a plus for the reader; particularly for those who still have mixed emotions concerning rectangular, circular or square waveguide; let your fears be put to rest. Included in the appendix of this report is a typical specification that any critical consultant engineer would want to include in his bid request package as applicable to waveguide systems of any type. A specification of this nature can only provide peace of mind and protect the welfare of his client over the useful life of the equipment package under consideration.

Richard E. Fiore

Mark A. Aitken

APPENDIX

LIST OF REFERENCE ARTICLES AND TEXT BOOKS

1. Microwave Transmission Design Data By Theodore Moreno. Dover Publications, 920 Broadway, N.Y. 10 N.Y. Dept. TF1.

2. Reference Data For Radio Engineers, Howard W. Sams & Co.

3. Fight Waveguide Losses 5 Ways, Richard M. Cox, Microwaves, August 1966, Page 32.

4. Mode Conversion In Circular Waveguides, E.R. Nagelberg and J. Shefer, Bell System Tech. Journal, September 1965, Page 1321.

5. "A Very High Power Variable Attenuator", R. Fiore, 1960, Published as RCA Technical Report TR-60-570-6

6. Andrew Corporation, Circular Waveguide Electrical Characteristics Product Catalog Page 80.

7. State Of The Waveguide Art, Tore N. Anderson, Microwave Journal, December 1982, Page 22.

APPENDIX

Comparison of Attenuation Characteristics of Circular to Square Waveguide

Assumptions

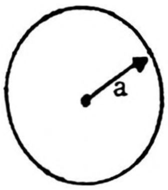
(a) Assume design frequency 635 MHz (average frequency of usable UHF high power television band 470 to 800 MHz).

(b) Assume waveguide dimensions based upon 635 MHz being the mid-frequency between the low frequency of cut-off and the high frequency which allows the propagation of the next higher order mode.

(c) Assume pure copper conductor

Circular Waveguide

The theoretical bandwidth of a circular waveguide operating at 635 MHz is:



$$a'' = 0.347\,\lambda_o$$

$$\lambda_o'' = \frac{11{,}800}{f_o\ (\mathrm{MHz})}\text{, inches}$$

$$a'' = 6.46''\ (\text{radius})$$

$$D'' = 12.92''$$

The expression for the theoretical attenuation of a copper circular waveguide of radius a'' is:

$$\alpha_c^{o} = \frac{0.423}{a^{3/2}}\left[\frac{\left(\frac{fo}{fco}\right)^{-\frac{1}{2}} + \frac{1}{2.38}\left(\frac{fo}{fco}\right)^{3/2}}{\sqrt{\left(\frac{fo}{fco}\right)^{2}-1}}\right]\ \mathrm{dB}/100'$$

a = Radius in inches

fo = 635MHz; fco = 550 MHz

$$\frac{fo}{fco} = \frac{635}{550} = 1.15$$

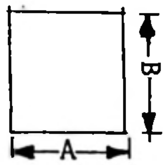
Thus $\left(\frac{fo}{fco}\right)^{-\frac{1}{2}} = (1.15)^{-\frac{1}{2}} = 0.933$

$$\left(\frac{fo}{fco}\right)^{3/2} = (1.15)^{3/2} = 1{,}23$$

$$\sqrt{\left(\frac{fo}{fco}\right)^{2}-1} = \sqrt{(1.15)^2-1} = 0.568$$

Square Waveguide

The theoretical bandwidth of a square waveguide operating at 635 MHz is:



A = B

$$A'' = \frac{\lambda_o''}{1.5}$$

$$\lambda_o'' = \frac{11{,}800}{1.5\ fo\ (\mathrm{MHz})}\text{, inches}$$

$$A = 12.39''$$

The expression for the theoretical attenuation of copper square waveguide of dimension A'' is:

$$\alpha_c^{\square} = \frac{1.107}{A^{3/2}}\left[\frac{\left(\frac{fo}{fco}\right)^{-\frac{1}{2}} + \frac{1}{2}\left(\frac{fo}{fco}\right)^{3/2}}{\sqrt{\left(\frac{fo}{fco}\right)^{2}-1}}\right]\ \mathrm{dB}/100'$$

A = Height or width in inches

fo = 635 MHz; fco = 476 MHz

$$\frac{fo}{fco} = \frac{635}{476} = 1.334$$

Thus $\left(\frac{fo}{fco}\right)^{-\frac{1}{2}} = (1.334)^{-\frac{1}{2}} = 0.866$

$$\left(\frac{fo}{fco}\right)^{3/2} = (1.334)^{3/2} = 1.541$$

$$\sqrt{\left(\frac{fo}{fco}\right)^{2}-1} = \sqrt{1.78-1} = 0.883$$

Substituting the derived values

$$c = \frac{0.423}{(6.46)^{3/2}} \left[\frac{0.933 + 0.518}{0.5679}\right] \text{dB}/100'$$

$$= 0.0258\ (2.555)$$

$$= 0.066\ \text{dB}/100'$$

Substituting the derived values

$$sq = \frac{1.107}{(12.39)^{3/2}} \left[\frac{0.866 + 0.77}{0.883}\right] \text{dB}/100'$$

$$= 0.0254\ [1.853]$$

$$= 0.047\ \text{dB}/100'$$

The conclusion that can be made from the attenuation calculations is that for a square waveguide system operating at the mid frequency between the waveguide cut-off frequency and that frequency at which the next higher order mode will be supported, the attenuation of square waveguide is 29% less than that of equivalent size circular waveguide.

ANALYTICAL AND EXPERIMENTAL DATA

ON CIRCULAR WAVEGUIDE FOR UHF-TV

A. G. Holtum
T. H. Greenway
T. J. Vaughan

MICRO COMMUNICATIONS, INC.

MANCHESTER, NEW HAMPSHIRE

1.0 INTRODUCTION

Circular waveguide transports energy and intelligence with less attenuation than other uniform cross-sections such as rectangular or square.

The dominant TE 11 mode in circular waveguide has been used for many years at microwave frequencies in the waveguides that connect the transmitter-receiver system to the antenna in many common carrier microwave radio relay links.

Micro Communications, Inc. has recently introduced circular waveguide for UHF-TV applications. In addition to the advantage of lower attenuation for a single channel, the circular guide can be used for transmitting two channels in orthogonal polarized modes, so that a single line can be used to feed two TV antennas.

The problems at UHF are the same as those for the microwave frequencies. That is, in using circular guide, we are mostly concerned with maintaining mode purity and preventing inadvertant conversion to higher order modes and/or inadvertant polarization coupling.

MCI has done considerable development work on the degree of mode purity and the amount of cross-polarization permitted. Measurements show that the performance can be optimized and the total characteristics determined from the input. A number of installations (up to 1500 feet) have been installed and tested.

There is a significant cost advantage when using circular waveguide with present high power UHF transmitters. For example, the transmission line cost for a 110 kW transmitter feeding a 1,500 foot tower (at channel 30) would be:

Coax, 8-3/16":	$282,700
Circular Waveguide:	$194,200

In addition to the initial hardware savings, there is a significant power savings due to the lower attenuation. The power at the far end delivered to the antenna would be:

Coax, 8-3/16":	79.6 kW
Circular Waveguide:	96.5 kW

With power costing about $1500 per kilowatt year, this 16.9 kW savings represents a $25,300 per year power bill savings.

2.0 NEED FOR CIRCULAR WAVEGUIDE

Circular and rectangular waveguides, unlike coax line, have no inner conductor. These types of transmission line are, in effect, high pass filters; that is, no energy will propagate below a critical frequency. Coaxial line, on the other hand, can conduct energy from very low frequencies to very high frequencies. The usable upper frequency for coaxial line is determined by the onset of higher order moding.

Power Handling

The average power handling capability of coaxial line depends on the allowable temperature rise of the inner conductor. The EIA specifies the maximum safe inner conductor temperature to be 100°C . With the increase in power of UHF transmitters from 55 kW to 110 kW and 220 kW, it is necessary to increase the size of the coax line from 6-1/8" to 8-3/16" and 9-3/16".

The recommended upper frequency and average power limits for coax line and waveguide are shown below:

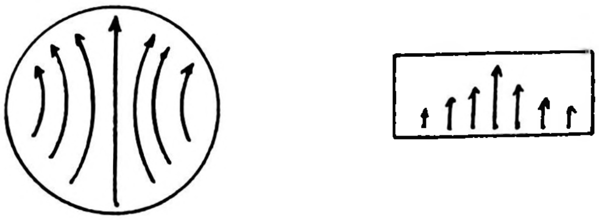
TRANSMISSION LINE SIZE	UPPER FREQUENCY LIMIT	UPPER POWER LIMIT
COAX:		
6-1/8 - 50Ω	816 MHz	56 kW
8-3/16 - 50Ω	613 MHz	120 kW
8-3/16 - 75Ω	689 MHz	80 kW
9-3/16 - 50Ω	543 MHz	180 kW
9-3/16 - 75Ω	611 MHz	120 kW
RECTANGULAR W/G		
WR-1800	640 MHz	1,000 kW
WR-1500	750 MHz	600 kW
WR-1150	960 MHz	350 kW
CIRCULAR W/G		
WC-1800	600 MHz	1,250 kW
WC-1500	725 MHz	750 kW
WC-1200	901 MHz	430 kW

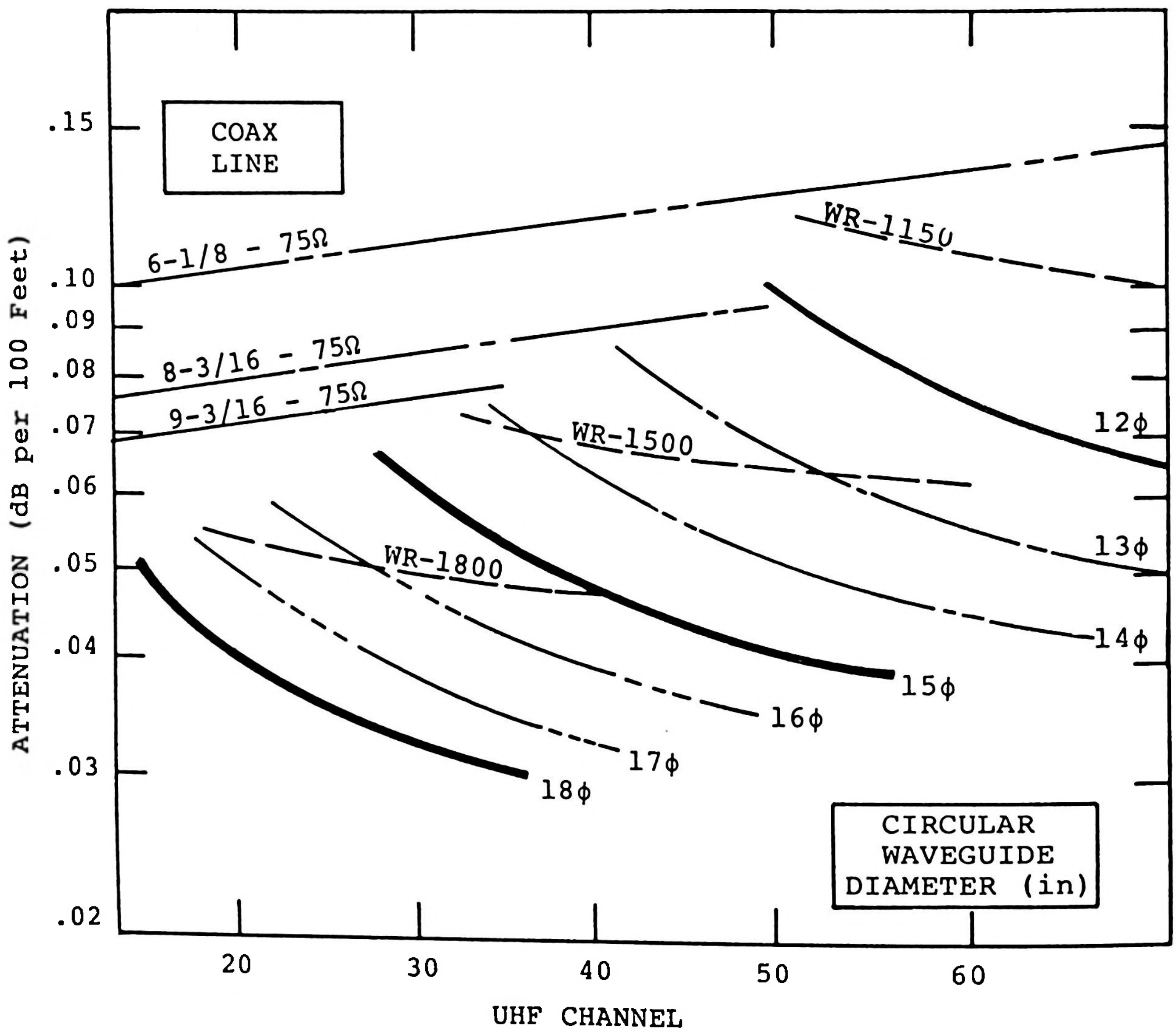
Efficiency

The efficiency of circular guide is significantly better than coax line and rectangular waveguide. The attenuation of the three types of lines is shown in Figure 1.

Moding

The electric field vectors for dominant mode propagation in rectangular and circular guide are shown below:

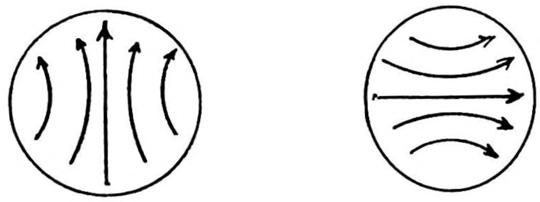




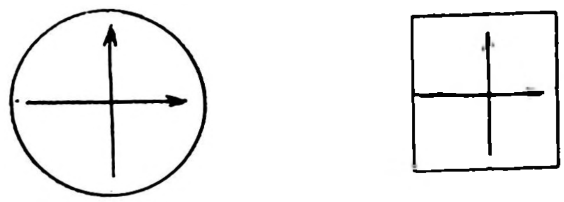
TRANSMISSION LINE ATTENUATION

FIGURE 1

Because the circular guide has the same cross-sectional dimension in two perpendicular planes, it is possible to propagate two independent fields (or 2 channels) in one guide.



This is possible with rectangular guide when the narrow dimension is made equal to the broad dimension. The resultant square guide, like the circular guide, can propagate two independent fields (or channels).



Both the rectangular and circular guide must have dimensions such that the operating frequency is well above the cut-off frequency.[1]

Moding problems will occur for either waveguide if too large a size is used. The EIA recommends the following:

	Rectangular	Circular
Lower Limit:	.62 fc TE 02	1.15 fc TE 11
Upper Limit:	.95 fc TE 02	.95 fc TE 21

---

1 See "MCI Wavelength Tables" and "Waveguide or Coax for 5 MW" by T. J. Vaughan, IEEE Transaction on Broadcasting, Vol. BC-16, March 1970

These criteria would result in the following channel designations:

Rect. W/G	-	Chan.	Cir. W/G	-	Chan.
WR-1800		14-39	WC-1800		14-35
WR-1500		17-60	WC-1500		24-55
WR-1150		42-70	WC-1200		46-70

It is obvious from Figure 1 that for circular waveguide, the larger the diameter, the higher the efficiency, since the attenuation for a fixed diameter continues to decrease with increase in frequency for the ranges considered.

$$\text{Efficiency (\%)} = 100 \times 10^{-\left(\frac{\alpha}{10}\right)}$$

The higher order modes that can propagate if excited in circular waveguide are shown in Figure 2. Energy cannot be converted to these modes if (a) they are not excited, and (b) if they are excited but the frequency is below the cut-off for that mode.

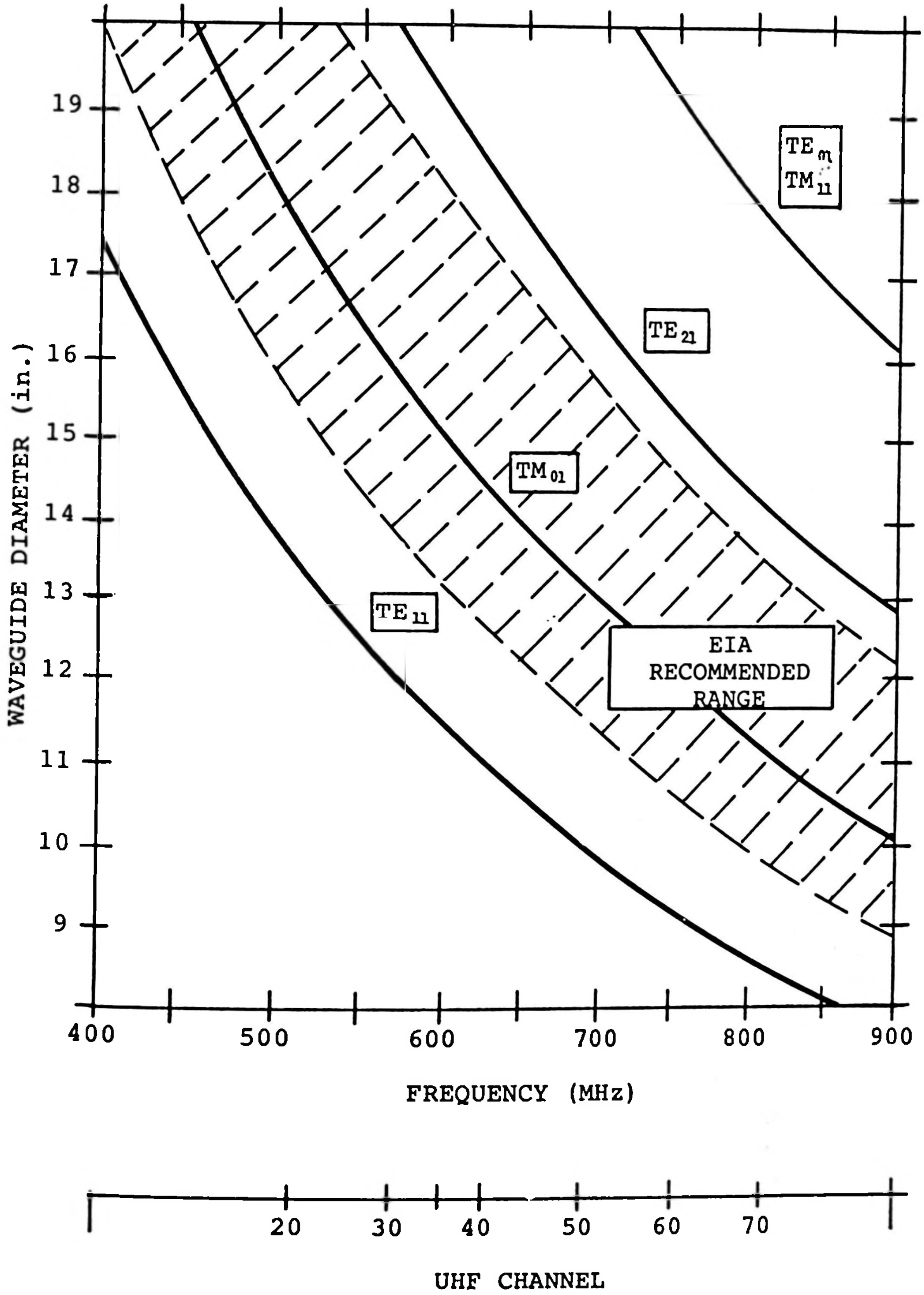
To insure that the higher modes are not excited in UHF-TV installations, rectangular waveguide is used in the horizontal run from the building to the tower, and rectangular elbows are used at the base of the tower. The transmission line is transformed from rectangular to circular* at the start of the vertical run and converted back again to rectangular waveguide at the top of the tower. Circular waveguide is therefore utilized only where it is most useful from attenuation and wind load points of view.

Polarization Rotation

Polarization rotation can occur in circular guide. This is not possible in the rectangular guide used for the horizontal run because the rectangular shape tends to polarize the field, and all cross-polarized fields, if excited, are below cut-off. The polarization rotation is corrected after installation by introducing, at the start of the long vertical run of circular waveguide, a complementary elliptical polarized wave.* The technique used results in a pure linearly polarized wave at the output end of the circular run with exceptionally low VSWR and high efficiency.

---

* Patent Pending



CIRCULAR WAVEGUIDE MODE CHART

FIGURE 2

3.0 THEORETICAL CONSIDERATIONS

The cut-off frequency for the first higher order mode, TM 01, in circular waveguide, is only 1.32 times the cut-off frequency of the dominant TE 11 mode, somewhat less than that for rectangular guide. To take full advantage of the lower attenuation afforded by the circular guide, we must operate in the region where the guide will support the TM 01 mode yet below the cut-off frequency of the TE 21, so that if this or higher order modes are excited, they will be attenuated in relatively short distances (see Figure 2).

Mode Conversion

Energy conversion to the higher order modes can occur at the rectangular to circular tranducer and at the flanges. The MCI transducer has been carefully designed so that no higher modes are generated.

Discontinuities due to flange misalignment while operating in the TE 11 mode can excite different spurious modes. The three basic types of discontinuities and the corresponding (power) coupling coefficients are shown in Table 1.

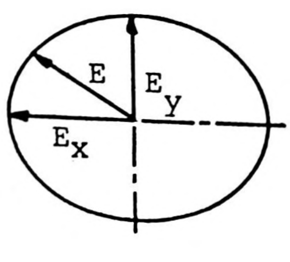
The spurious modes generated are well below cut-off for the recommended diameters. The tolerance to diameter ratio of circular guide is about 1000:1, therefore coupling to these higher modes will be down approximately 60 dB.

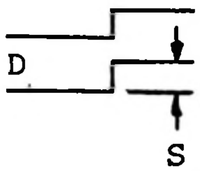
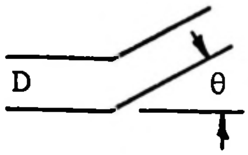
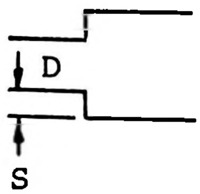
Cross Polarization

A problem area in any waveguide capable of supporting two orthogonal signals in the dominant mode is the cross-coupling effect (even if only one polarization is used). Cross-coupling is caused by asymmetry of the guide cross section such as ellipticity.

Elliptical guide can cause cross-coupling if the direction of the E-vector is not coincidental with the major or minor axis of the ellipse.

Shown below is an exaggerated elliptical cross section and the relative position of the incident E-vector.



DISCOUNTINUITY	SPURIOUS MODE	POWER COUPLING
Off-Set: 	TE 01	$\left[2\sqrt{2}\left(\frac{S}{D}\right) K\right]^2$
Tilt: 	TE 01	$\left[2\sqrt{2}\pi^2 \left(\frac{D}{\lambda}\right)\left(\frac{\theta}{\pi}\right) K\right]^2$
Diameter Change: 	TM 11	$\left[4 \left(\frac{S}{D}\right)\right]^2$

Ref: "Low Loss Transmission Using Overmoded Waveguide" by T. Anderson, IEEE-PGAP/MTT - 1981.

$$K = \frac{K\ 01\ (K\ 11)^2}{(K\ 01^2 - K\ 11^2)(K\ 11^2 - 1)^{\frac{1}{2}}}$$

where $K\ 01 = 3.832$

$K\ 11 = 1.841$

TABLE 1

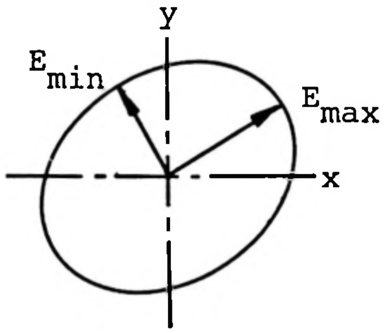
TE 11 TO SPURIOUS MODE COUPLING DUE TO CIRCULAR WAVEGUIDE DISCONTINUITIES

An initially linearly polarized wave can be broken into two orthogonal components, Ex and Ey, parallel to the major and minor axes as shown. Because of the geometry, the two components travel at different velocities with the result that they are out of phase by an amount:

$$\Psi = 2\pi\ell\left(\frac{1}{\lambda_{gx}} - \frac{1}{\lambda_{gy}}\right)$$

where λ_{gx} and λ_{gy} are the wavelengths in the guide for the respective components and ℓ is the length. The wave is now elliptically polarized. The degree of cross-polarization may be defined by the axial ratio, (AR). Expressed in dB, the axial ratio is:

$$AR = 20 \log\left[\frac{E_{max}}{E_{min}}\right]$$



Because of the randomness of perturbations such as these, the major axis of the polarization ellipse will also be rotated by a small angle relative to the incident polarization. The most severe axial ratio is obtained when the angle of the electric vector is oriented 45 degrees with respect to the major axis of the guide ellipse.

Recombination of Orthogonal Polarizations

It is not possible to manufacture circular waveguide sections with tolerances of sufficient precision that its ellipticity will be insignificant nor is it feasible to insure that the electric vector will always be coincidental with the major or minor axis. Fortunately, if a component is generated which is orthogonal to the incident polarization, it can be corrected by generating another component which, when combined with the original, will produce a linearly polarized wave at the far end.

For very long runs of circular guide, we have only to make a single correction in order that, when the elliptically polarized wave arrives at the far end transducer, it is linearly polarized. This is accomplished by introducing a complementary ellipse at the tower base. In almost all cases, virtually complete compensation is achieved, with the result that linear polarization is restored, though the plane of polarization may be rotated somewhat from that of the original incident wave. This, in turn, is corrected by rotating the output transition at the far end of the waveguide run at the top of the tower.

In the case where dual polarized signals are used, this technique automatically compensates both polarizations. The technique depends only on the geometry of the cross section and is relatively insensitive to temperature and frequency.

4.0 MEASUREMENTS

Measurements made on circular waveguide of runs up to 1500 feet show that:

(a) Using a special transducer for the TE 11 mode, higher order modes will not be excited.

(b) The cross polarized component normally excited can be removed so that there is a pure linearly polarized wave at the far end.

(c) The total attenuation due to any residual cross polarized component can be determined from Return Loss measurements.

(d) Return Loss (VSWR) of circular waveguide runs are much better than rectangular or square waveguides.

By means of both reflection and attenuation measurements, it was determined that with the new TE 11 mode transducer, no higher modes were excited. The cross-polarized component of the TE 11 modes which does exist and results in an elliptically polarized wave at the far end is correctable.

The complete properties of any transmission line can be determined by attenuation and reflection measurements. The reflection measurement was made using a precision bridge and and a Spectrum Analyzer. Return Loss was measured over the desired frequency range under many conditions with (a) TE 11 mode cross-polarized component present and (b) with the TE 11 mode cross-polarized component removed. Attenuation was measured under identical conditions using the cavity resonance method.

The results are shown in Figure 3A and 3B. It is clearly evident from Figure 3A that, although the major axis of the TE 11 wave is aligned with the far end transition, the wave is elliptically polarized. Most of the energy is coupled to the output transition. Some, however, is reflected and the transmission line acts like a high Q resonant cavity for the cross-polarized component.

The corresponding attenuation measurement shows there is a relationship between the level of the spikes and the loss at that frequency. The total attenuation is:

$$\alpha_T = \alpha_{cross} + \alpha_{normal}$$

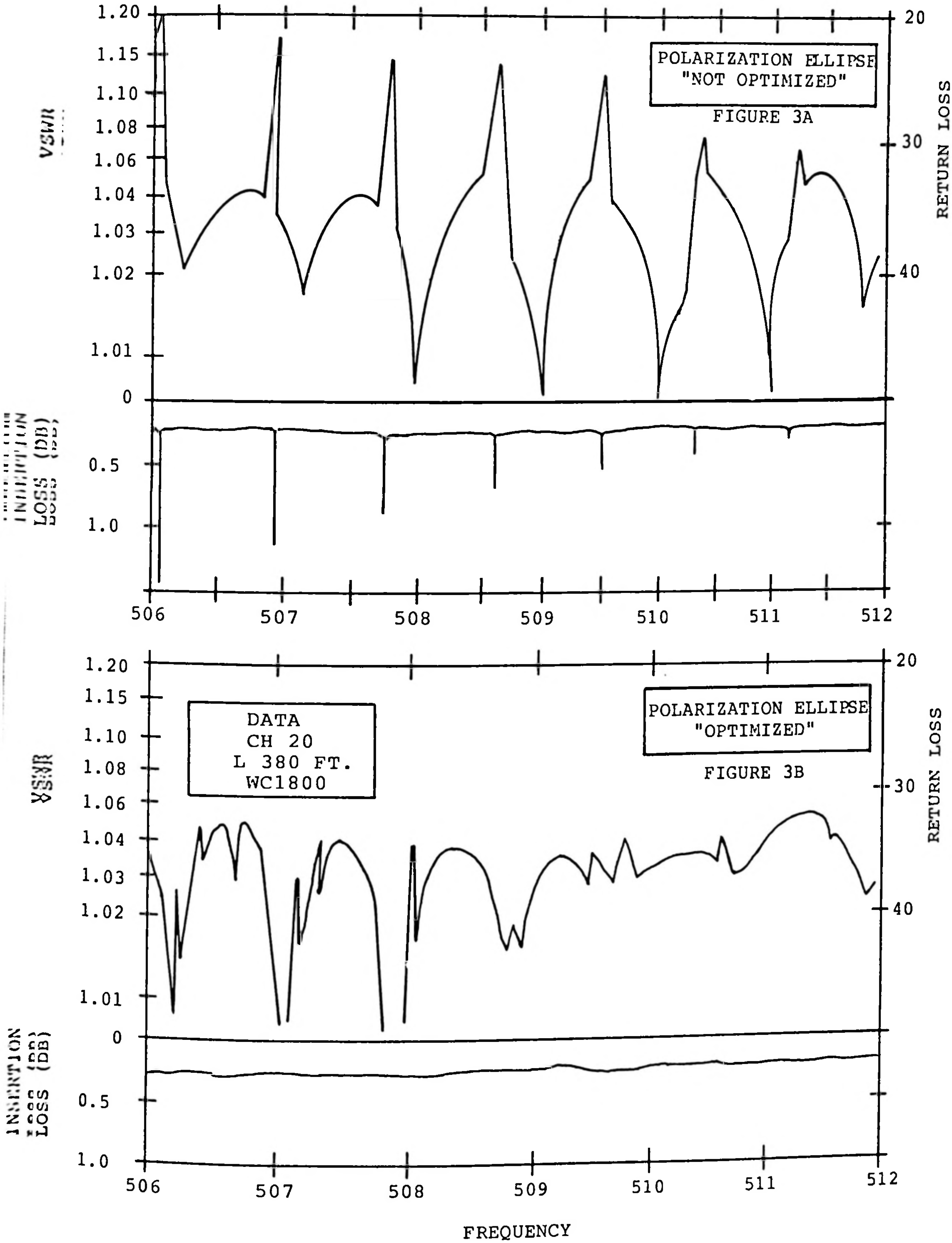
The loss due to mismatch is neglected since it is only .004 dB for a Return Loss of 30 dB (VSWR 1.06). Total insertion loss was measured for a large number of cross-polarization spikes which were deliberately introduced. The results are plotted against spike Return Loss (Figure 4). The data shows that for the spike loss to be negligible, the spike Return Loss should be less than (a larger number than)

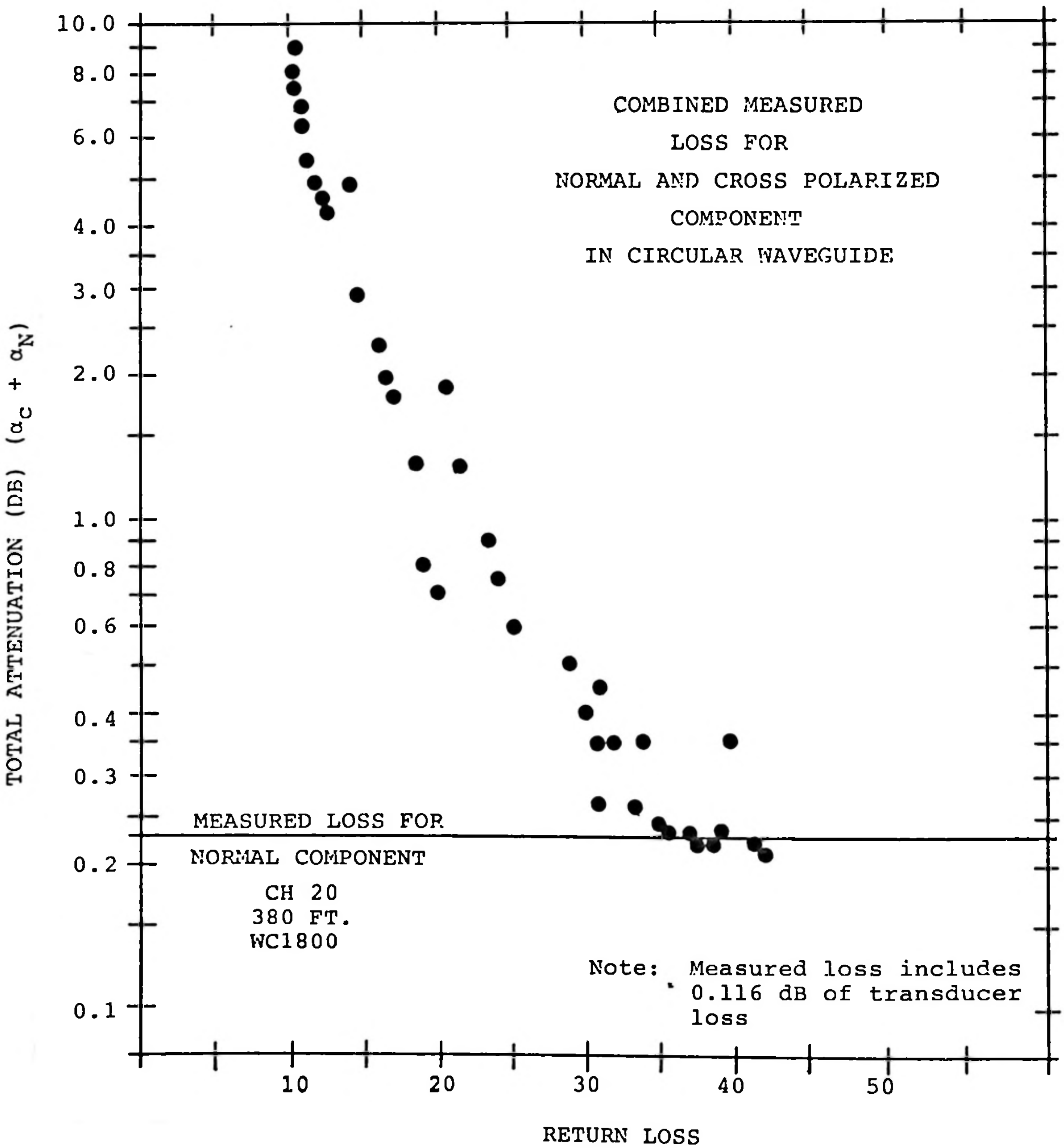
Return Loss		Typical Run
24 dB	-	1500 Feet
26 dB	-	1000 Feet
30 dB	-	380 Feet

and that, when the cross polarized component is removed, the attenuation measured is the calculated value for the guide in question.

After each test where the cross-polarization was deliberately introduced, the same waveguide was then optimized by compensation. As shown in Figure 3B the spikes disappeared and the measured insertion loss was then equal to the theoretical I^2R loss in the guide. The test transition used had been previously measured to have a loss of 0.116 dB with an rms error of +/- 0.01 dB.

The VSWR or Return Loss of a long circular waveguide run tends to be much better than that of rectangular waveguide because the impedance-determining dimension is twice that of rectangular waveguide and the tolerance to diameter ratio is 1000:1 as compared to 250:1 for rectangular guide. Flange discontinuity of circular guide is negligible. VSWR measurements made on many installations, some up to 1500 feet, show no flange build-up, unlike rectangular waveguide or coax line.


1.20
1.15
1.10
1.08
1.06
1.04
1.03
1.02
1.01
0
VSWR
0.5
1.0
INSERTION LOSS (DB)
20
30
40
RETURN LOSS
POLARIZATION ELLIPSE
"NOT OPTIMIZED"
FIGURE 3A
506
507
508
509
510
511
512
1.20
1.15
1.10
1.08
1.06
1.04
1.03
1.02
1.01
0
VSWR
0.5
1.0
INSERTION LOSS (DB)
20
30
40
RETURN LOSS
DATA
CH 20
L 380 FT.
WC1800
POLARIZATION ELLIPSE
"OPTIMIZED"
FIGURE 3B
506
507
508
509
510
511
512
FREQUENCY



COMBINED MEASURED
LOSS FOR
NORMAL AND CROSS POLARIZED
COMPONENT
IN CIRCULAR WAVEGUIDE
TOTAL ATTENUATION (DB) $(\alpha_C + \alpha_N)$
10.0
8.0
6.0
4.0
3.0
2.0
1.0
0.8
0.6
0.4
0.3
0.2
0.1
MEASURED LOSS FOR
NORMAL COMPONENT
CH 20
380 FT.
WC1800
Note: Measured loss includes 0.116 dB of transducer loss
10
20
30
40
50
RETURN LOSS

FIGURE 4

A plot of VSWR and measured insertion loss for a typical 1000 foot installation is shown in Figure 5.

5.0 DUAL CHANNEL CIRCULAR WAVEGUIDE

Many UHF installations have more than one antenna at the tower top. Most of the new waveguide installations are mounted inside the tower in space normally occupied by other lines, ladders, elevators, etc. A dual channel circular waveguide will not only have a significant wind load advantage, but will also save space.

The design problems associated with this type of transmission line are:

(a) Band width at each channel

(b) Coupling between the two transmitter at the tower base

(c) Coupling into the other antenna

The band width is the principle design problem. The dual mode tranducers must operate over the band width of both channels. In any installations, this is a minimum of six channels or an 8% band width as compared with a 1% band width for a single channel application.

It is necessary to maintain a 30-35 dB isolation between the transmitters at the tower base to reduce intermodulation. It is also desirable to maintain at least a 35 dB isolation between antennas to minimize ghosting problems.

Measurements made on a dual channel system (Figure 6) show that this performance can be achieved.

TYPICAL CIRCULAR

WAVEGUIDE INSTALLATION

CH 28

TOWER 964 FT.

HOZ. RUN 78 FT.

CALCULATED ATTENUATION	α	MEASURED ATTENUATION FREQUENCY	α
Elbows/Twists	0.040	555.25	0.550
Transducer #1	0.055	555.80	0.610
Transducer #2	0.055	556.20	0.610
Rectangular Guide(78')	0.053	556.6	0.696
Circular Guide(964')	0.400	557.2	0.610
NET (DB)	0.603	558.2	0.555
		558.5	0.610
		AVG. (DB)	0.605

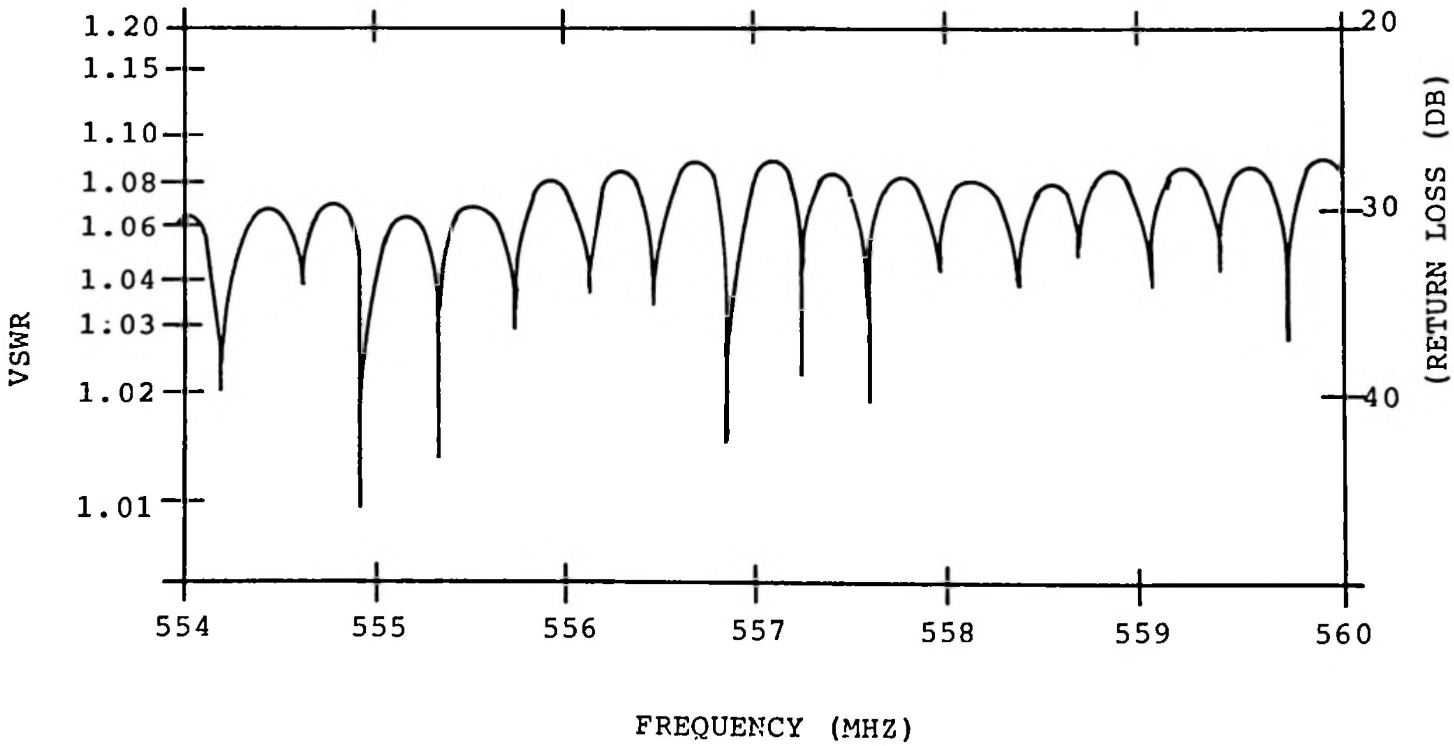


FIGURE 5

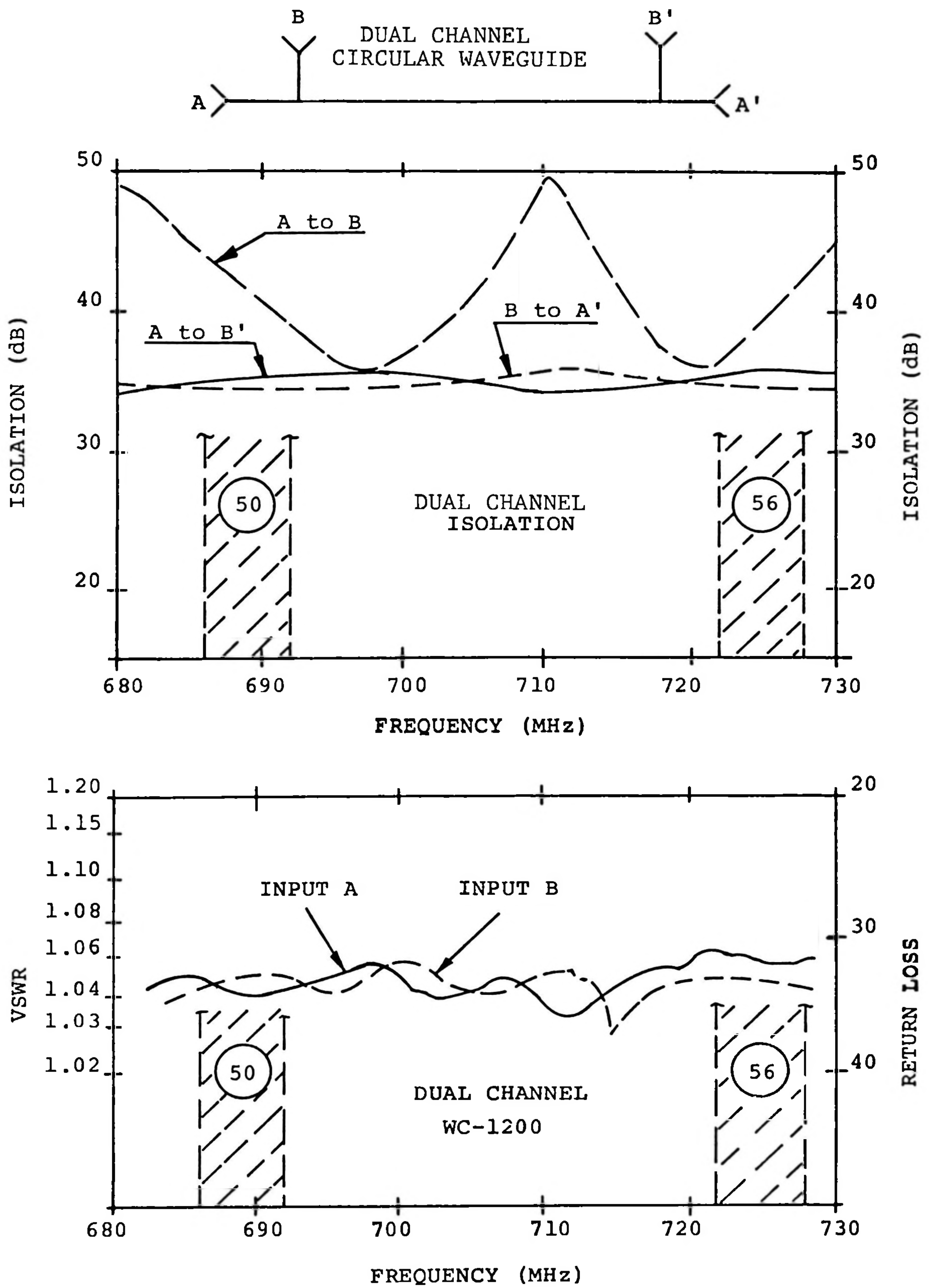

B
B'
DUAL CHANNEL
CIRCULAR WAVEGUIDE
A
A'
50
40
30
20
50
40
30
20
ISOLATION (dB)
ISOLATION (dB)
A to B
B to A'
A to B'
50
56
DUAL CHANNEL
ISOLATION
680
690
700
710
720
730
FREQUENCY (MHz)
1.20
1.15
1.10
1.08
1.06
1.04
1.03
1.02
20
30
40
VSWR
RETURN LOSS
INPUT A
INPUT B
50
56
DUAL CHANNEL
WC-1200
680
690
700
710
720
730
FREQUENCY (MHz)

FIGURE 6

COMPARISON OF C BAND AND Ku BAND FOR BROADCAST QUALITY TELEVISION TRANSMISSION BY SATELLITE

Norman P. Weinhouse
Hughes Communications, Inc.

ABSTRACT

This paper compares C band and Ku band satellite transmission for delivery and distribution in the United States of broadcast quality television signals. Both existing and proposed satellites are considered. The standard of comparison is EIA document RS250B, which calls for a signal-to-noise (S/N) ratio of 56 dB for satellite transmission and a 99.99 percent continuity of service. The satellite link is quantified in terms of satellite power (EIRP) and receiving station merit (gain to temperature ratio). Factors which affect reliability of propagation are discussed and are quantified for the diverse weather conditions in the United States. Results of the comparison show a slight clear weather advantage to Ku band but an overwhelming advantage to C band to achieve coninuity of service objectives. Overcoming interference from terrestrial microwave sources at C band is discussed.

INTRODUCTION

This document presents an objective comparison of the use of Ku band (12 GHz) and C band (4 GHz) for satellite delivery of "high quality" television for broadcast service. The standard used is EIA document RS250B, specifically generated for transmission of broadcast television. The parameters assumed for this comparison are signal-to-noise ratio (paragraphs 5.3.1 and 6.3 of RS250B) and continuity of service (paragraphs 5.4 and 6.5 of RS250B) since these are the only factors affected by propagation. All of the other parameters are related to system equipment, which can be selected to support the requirements regardless of the radio frequency used. It is assumed that frequency modulation is used in the satellite link.

SATELLITE LINK

Figure 1 graphically depicts a satellite link. For simplicity the satellite is shown to have gain and to contain a noise generator. The link consists of two microwave hops called uplink and downlink.

The overall predetection carrier-to-noise (C/N) ratio is

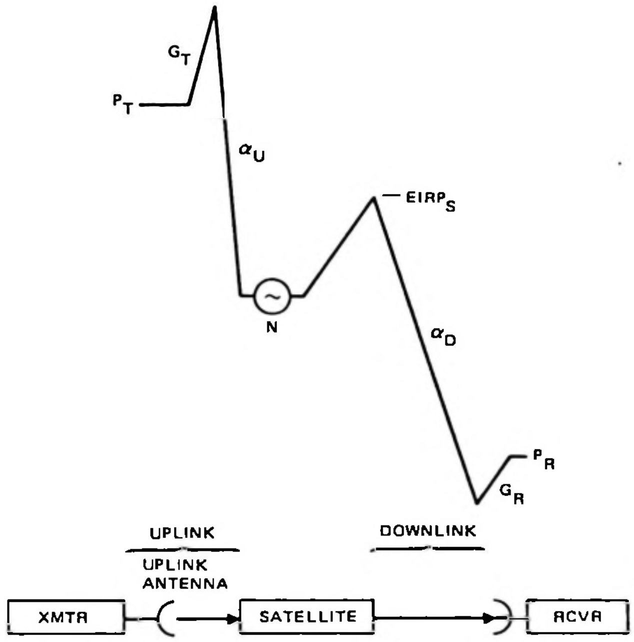


FIGURE 1. SATELLITE LINK POWER CHART

$$(C/N)_{PD} = (C/N)_U \oplus (C/N)_D \oplus (C/I)_T \qquad (1)$$

where

$\oplus$ denotes power summation

$(C/I)_T$ is total carrier-to-interference power ratio

$$(C/N)_U = [\,P_T + G_{AT} - \alpha_U + (G/T)_S - (k + B)\,]\ \text{dB}$$

$$(C/N)_D = [\,(EIRP)_S - \alpha_D + (G/T)_R - (k + B)\,]\ \text{dB}$$

For an FM receiver operating above threshold, the video signal-to-noise ratio (peak luminance signal to rms noise) can be expressed by

$$(S/N)_V = 6\ m^2\ \left(\frac{B}{fm}\right)\ (C/N)_{PD}\ (PW) \qquad (2)$$

where

m is modulation index $(\frac{\Delta F}{fm})$

ΔF is deviation of main carrier by video

B is predetection bandwidth

fm is highest modulation frequency (4.2 MHz for NTSC)

(pw) is combined pre-emphasis and weighting improvement factor (12.8 dB for U.S. and CCIR standards)

This equation can be expressed in decibels as

$$(S/N)_V = [10 \log \Delta F^2 + 10 \log B + (C/N)_{PD} - 178.1] \text{ dB}$$

The standard for this parameter from RS250B is 56 dB. Prudence and good design practice dictate that at least 2 dB margin be included to guard against equipment degradation and/or antenna pointing errors. Using $(S/N)_V = 58$ dB and rearranging terms, the clear weather $(C/N)_{PD}$ would be

$$(C/N)_{PD} = 236.1 - 10 \log (\Delta F^2 B) \text{ dB} \quad (3)$$

This expression is shown graphically in Figure 2 as a plot of $(C/N)_{PD}$ versus deviation ΔF, assuming a fixed relation between bandwidth (B) and ΔF as $B = 2(\Delta F+7)$. It is no suprise that for a given signal-to-noise ratio, wider deviation allows for a lower carrier-to-noise ratio. However, there is another important practical consideration shown in Figure 2. The very best threshold extension receivers exhibit impulse noise with a predetection carrier-to-noise of 8 dB or lower. Conventional receivers exhibit impulse noise at a predetection C/N of approximately 12 dB. This means that a system providing a given S/N using wide deviation would have less fade margin than the system using narrow deviation which provides the same S/N.

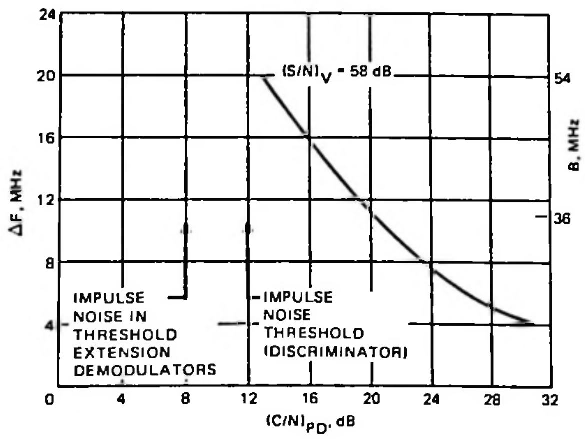


FIGURE 2. FM DEVIATION VERSUS C/N, SATELLITE LINK

C BAND LINK - CLEAR WEATHER

Current and projected C band (6 GHz up, 4 GHz down) U.S. domestic satellites have a frequency reuse plan such that each transponder has a 36 MHz bandwidth. One authorized satellite operator has some transponders with a 72 MHz bandwidth. All C band statellites are designed such that if a saturating signal is sent from the uplink station, the uplink clear weather C/N is greater than 30 dB. Furthermore, internal interference and external interference with well designed earth stations will produce an overall $(C/I)_T$ greater than 23 dB. From Equation 1, the predetection carrier-to-noise ratio is

$$(C/N)_{PD} = (C/N)_D \oplus 30 \oplus 23 \text{ dB}$$

$$= (C/N)_D \oplus 22.2 \text{ dB}$$

From above considerations $[(S/N)_V = 58 \text{ dB}]$, the

$$(C/N)_{PD} = 19.7 \text{ dB}$$

Therefore

$$(C/N)_D = 21.7 \text{ dB}$$

From the above expression for $(C/N)_D$, and substituting appropriate values

$$21.7 \text{ dB} = [\text{EIRP} - (196.0 \pm 0.5) + G/T + 228.6 - 75.6] \text{ dB}$$

$$(\text{EIRP} + G/T) = (64.7 \pm 0.5) \text{ dB} \quad (4)$$

The ±0.5 dB factor is due to the difference in slant range from the satellite to various points in the U.S., assuming the satellite is at the extremes of the assigned orbital arc.

KU BAND LINK - CLEAR WEATHER

The Ku band link is somewhat difficult to quantify since there has been limited experience. The existing and proposed transponders have bandwidths from 54 MHz to 36 MHz (the proposed direct broadcast system (DBS) standard is 18 MHz). There is a wide range of uplink performance as well as a wide range of EIRP for various kinds of coverage. Here, in all cases the best case conditions are assigned to the clear weather Ku band link.

1) Assume $(C/N)_U = 30$ dB

2) Assume $(C/I)_T = 30$ dB

Therefore

$$(C/N)_{PD} = (C/N)_D \oplus 27 \text{ dB}$$

With an available bandwidth of 54 MHz, the $(C/N)_{PD}$ can be 12.8 dB to produce a $(S/N)_V$ of 58 dB

Therefore

$$(C/N)_D = 13 \text{ dB}$$

Making the appropriate substitutions (as in the C band case above)

$$13 \text{ dB} = [\text{ EIRP} - (205.5 \pm 0.5) + G/T + 228.6 - 77.3]$$

$$(\text{EIRP} + G/T) = (67.2 \pm 0.5) \text{ dB} \qquad (5)$$

C BAND AND KU BAND COMPARISON - CLEAR WEATHER

An appropriate assumption for C band EIRP for CONUS coverage would be 35 dBW for current satellites. Again, giving clear weather benefit to Ku band with "time zone beams," an appropriate assumption for EIRP would be 48 dBW. A comparison of receiving earth station G/T for overall CONUS is:

G/T = 29.2 dB to 30.2 dB (C band)

G/T = 19.7 dB to 20.7 dB (Ku band)

Current technology indicates that the C band requirement in clear weather can be met with 10 meter diameter antennas with 80°K low noise amplifier (GaAsFET).

The Ku band requirement in clear weather could be met with 3 meter diameter antenna and 400°K low noise amplifier (LNA).

On the assumption that a large number of receiving stations (network affiliates) exist, and the uplink and satellite costs are essentially the same, the receiving station cost advantage is with the Ku band transmission under clear weather conditions. It should be pointed out, however, that if the Ku band satellite should have an EIRP of 43 dBW or less, or the combination of EIRP and/or bandwidth were to be reduced by a factor of 5 dB, there would be no economic advantage to Ku band even in clear weather.

PERFORMANCE WITH FADING

Paragraph 5.4 of RS250 is given here in its entirety:

5.4 Continuity of Video Service

Definition: The performance tolerances designed into a television relay system as detailed in this standard are not only system objectives describing the initial performance of such a system but also describe the performance of that system with time to the extent indicated in the standards that follow.

Standards:

a) The objective for continuity of video service is 99.99 percent of the annual operating time.

b) The continuity of video service will be deemed as interrupted if one or more of the following conditions exist.

(1) There is a total loss of the desired video signal.

(2) The peak-to-peak luminance level to rms weighted noise ratio is less than 37 dB.

(3) The performance standards are exceeded to the extent that the picture is not usable.

Method of Measurement: A chart recorder or other suitable device may be used to determine availability of the video signal. Parameters of pargraph b. provide guidelines for calculation of system statistical performance. Methods of continuous measurement of some parameters for real systems cannot be readily established at this time.

In a satellite transmission, total loss of signal will occur with equipment failure, a deep fade, or sun transit outage. The subject of equipment and sun transit is not considered here sinced this is a relative comparison of radio frequency band usage.

Futhermore, in a satellite transmission an "unusable picture" will occur before the S/N is less than 37 dB because impulse noise will be controlling and it occurs at C/N of about 8 dB in the very best receivers. This establishes the criteria for the faded link.

In general, satellite links need not provide for the same level of protection from fading as terrestrial links, since only a small portion of the path passes through the atmosphere. However, the space link is not entirely free from antmospheric effects and the attendant degradations to radio wave propagation. Satellite link propagation has been the subject of intense study and a firm data base has been established from experiment and theoretical studies. There are many degrading factors such as clouds, fog, turbulence, adiabatic effects, and precipitation. The effects of these factors are attenuation, depolarization, and sky noise increase. Secondary effects are scintillation (multipath), antenna gain degradation, and bandwidth coherence reduction due to variations of these effects with frequency. At C band, it has been demonstrated that in the contiguous 48 states, a 3 dB margin is all that is required for all atmospheric effects for more than 99.99 percent propagation reliability. Even in the worst geographical locations, worldwide a 6 dB margin will provide more than 99.99 percent propagation reliability. Above 10 GHz, however, the effects change dramatically, especially degradations due to precipitation.

Several models exist to predict attenuation due to rain[1]. The Crane Global Attenuation Model[2] is recommended for use in planning satellite systems because it is quantitative and has been proven accurate. The Global Model relates point rain rate

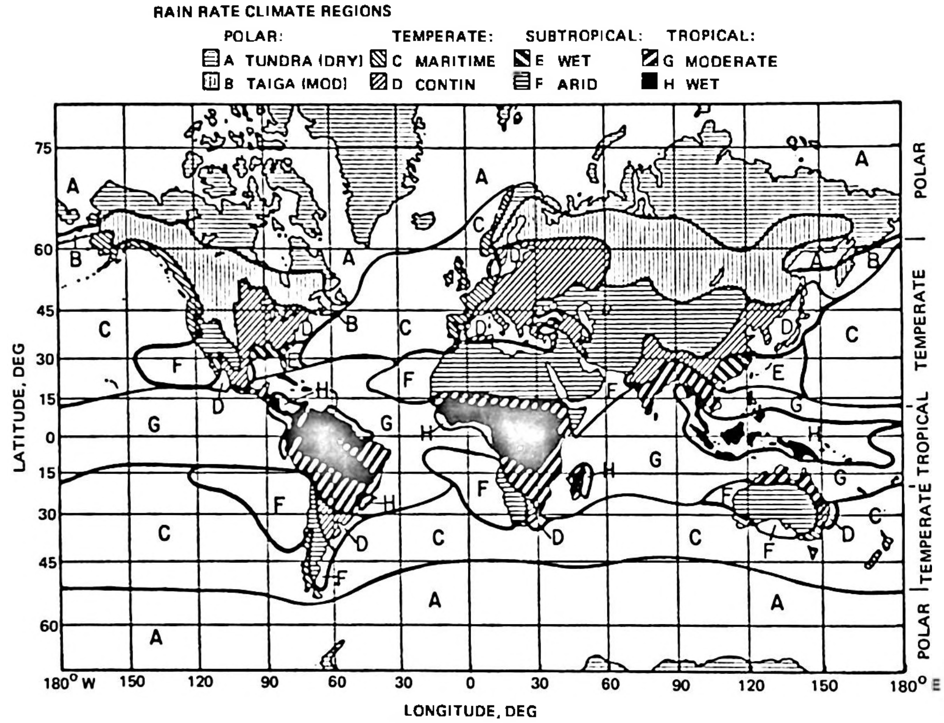


FIGURE 3. RAIN RATE CLIMATE REGIONS FOR THE GLOBAL RAIN ATTENUATION MODEL

(for which much statistical data exists) to attenuation at various frequencies. The Global Model correlates well with measured attenuation data[3]. Figures 3 and 4 show the rain rate climate regions worldwide, and for CONUS in particular. Figures 5 through 10 provide data for all of the rain rate regions in the CONUS. It is obvious that for nationwide network broadcast quality television distribution, a very wide range of fade margin must be provided for any decent percentage of propagation reliability at Ku band.

There are many interesting comparisons that can be made. A few are presented here.

1) Given the parameters established above for the clear weather comparison case, what is the continuity of service due to propagation?

Answer:

a) For C band: For all cases, very much greater than 99.99 percent.

b) For Ku band:

Best Case - for satellite position 100°W longitude, only Colorado and part of Wyoming

.... 99.99 percent

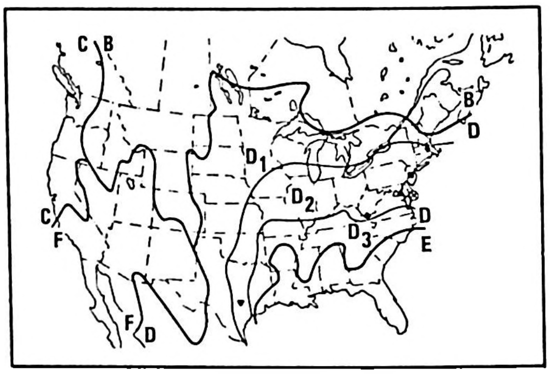


FIGURE 4. RAIN RATE CLIMATE REGIONS FOR THE CONTINENTAL U.S. SHOWING THE SUBDIVISION OF REGION D

Worst Case - (Rain Region E, satellite at 135°W)

.... 99.5 percent

Note: The difference between 99.5 percent and 99.99 percent is 43 hours per year.

2) Given today's technology, and adjusting the ground terminal parameters for equal cost between C and Ku band, and still maintaining clear weather signal-to-noise ratio objectives, what is the continuity of service due to propagation?

Approximate cost parity will exist with a Ku band antenna of 5.5 meter diameter versus a 10 meter diameter at C band, both with GaAsFET LNAs. Furthermore, the bandwidth and hence the deviation can be reduced in the Ku band transmission to enhance fade margin at the expense of clear weather signal-to-noise ratio. This is not much of a penalty since the noise floor of most receivers produces a S/N of about 62 dB. This all adds up to a 13 dB margin requirement at Ku band.

Answer:

a) For C band: For all cases, very much greater than 99.99 percent.

b) For Ku band:

Best Case - (Regions B, C, and F) .... 99.99 percent

Worst Case - (Region E) .... 99.90 percent

Note: The difference between 99.99 percent and 99.9 percent is 8 hours per year.

c) Assuming cost parity in the earth segment, what satellite EIRP at Ku band is needed to meet the 99.99 percent continuity of broadcast quality service objective at all points in the CONUS?

Answer:

a) Best satellite location .... 57 dBW

b) Satellite at edge of arc .... 67 dBW

In these comparisons, only precipitation attenuation effects were considered. A more complete analysis would include the following:

1) Sky noise contribution due to rain - Ippolito[3], in his excellent survey paper (63 references), quantifies the additional system noise due to rain fades. The Ku band receiver noise temperature in our comparison was assumed to be 460°K under clear sky conditions. With a 13 dB fade, the increase in system noise temperature is 254°K. This represents a degradation in system temperature of 1.9 dB. The system would therefore need nearly 15 dB margin to overcome a 13 dB fade.

2) Depolarization - In frequency reuse systems, satellite channels are staggered in frequency and adjacent channels are crosspolarized. Depolarization in the path will cause attenuation, but, more importantly, it will destroy the crosspolarization isolation in the receiving antenna. Some data exists, but an accurate model has not been established for prediction. Tests made at 12 GHz in Austin, Texas and Blacksburg, Virginia indicate for 0.01 percent of the time, crosspolarization isolation was in the range 20 to 21 dB. For a "broadcast quality" televison transmission, this amount of isolation would be marginal, depending on the information content in the adjacent crosspolarized transponders. If FM-TV were in the adjacent crosspolarization channels, approximately 50 percent of viewers would see the effects of interference. In less favorable rain locations, depolarization could give less than 99.9 percent continuity of service.

3) Uplink - It was assumed above that the uplink contribution is negligible. While this is reasonable at C band, it is giving the Ku band an advantage in the comparison. Ku band uplinks employ an adaptive power output control which not only adds to the cost, but these systems have response time which should be taken into account in a continuity of service analysis. Additionally, these adaptive systems are designed for about 10 dB correction. To overcome deep fades associated with high continuity of service objectives, additional complexity and very high cost amplifiers could be required.

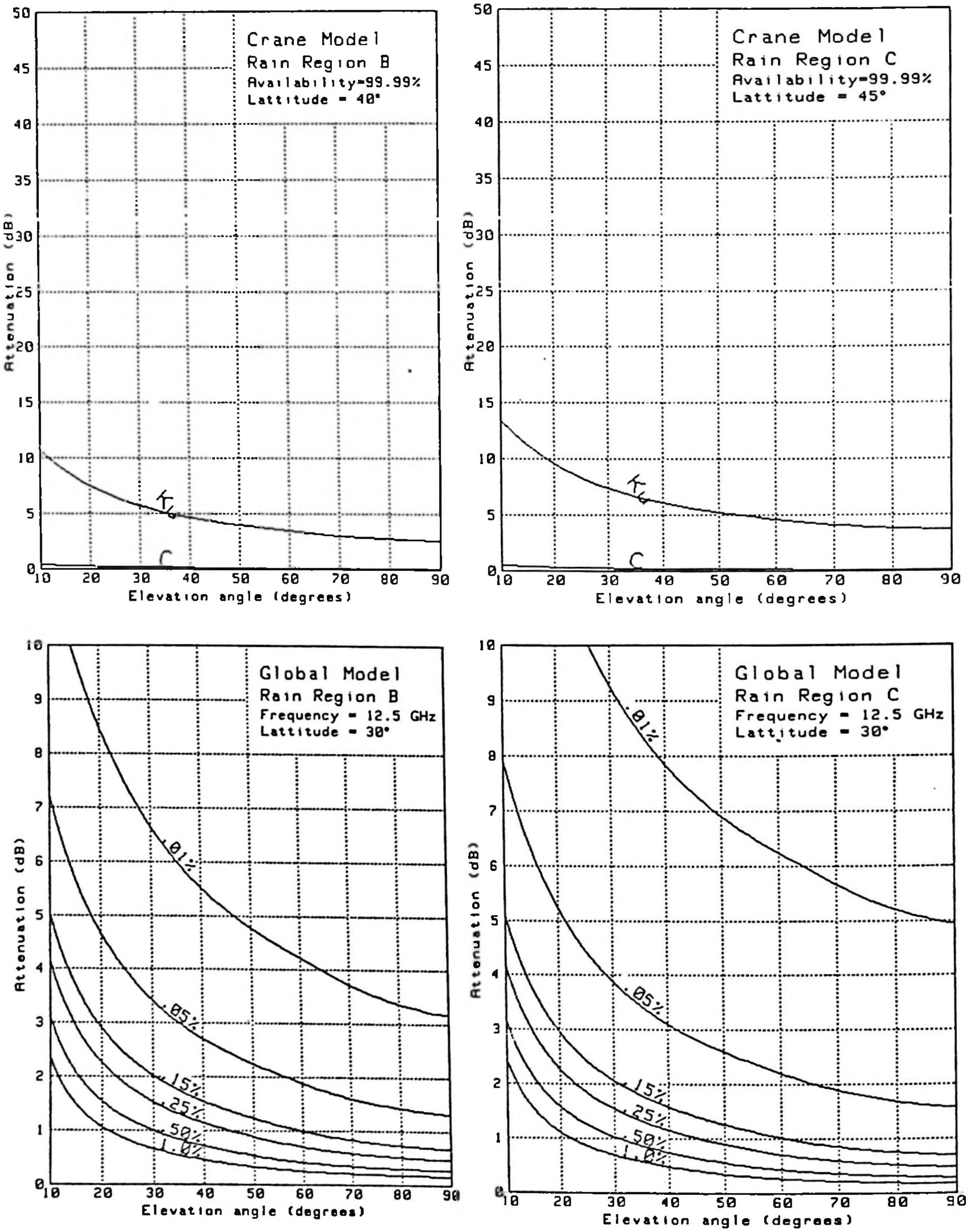


FIGURE 5. RELIABILITY DATA – RAIN RATE REGIONS B AND C

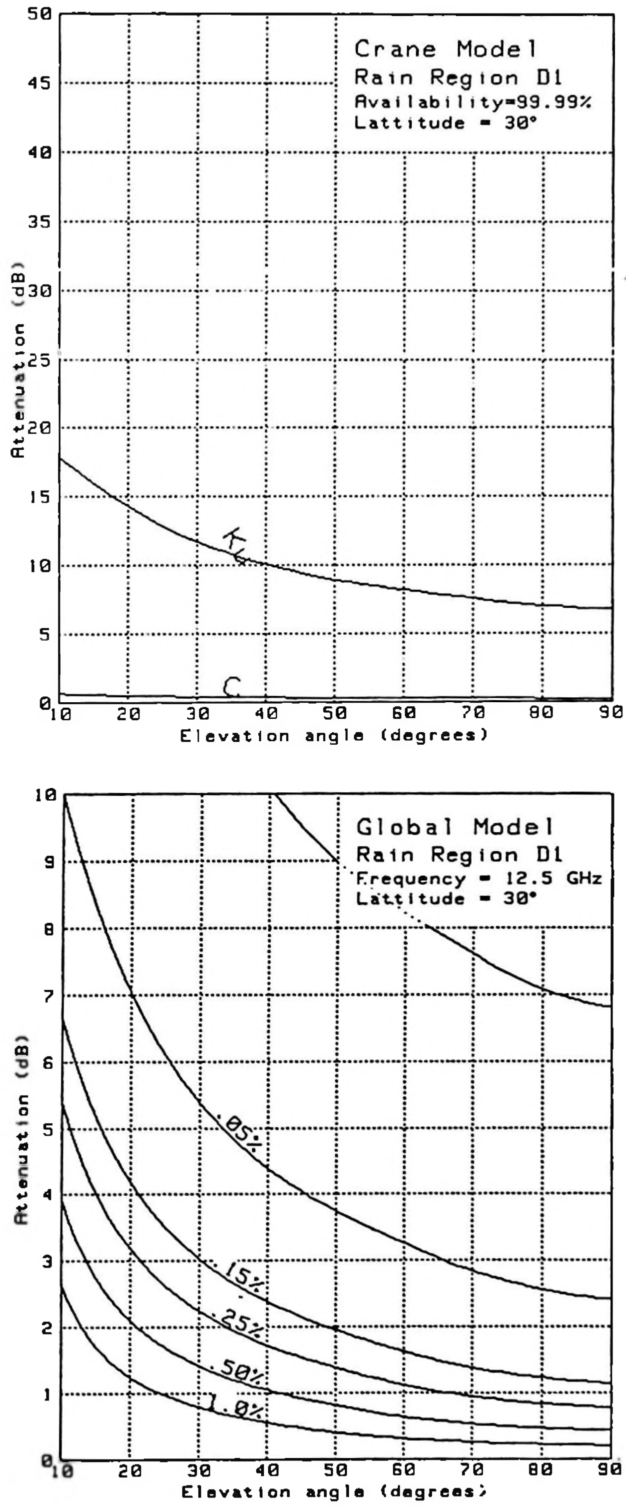


FIGURE 6. RELIABILITY DATA – RAIN RATE REGION D1

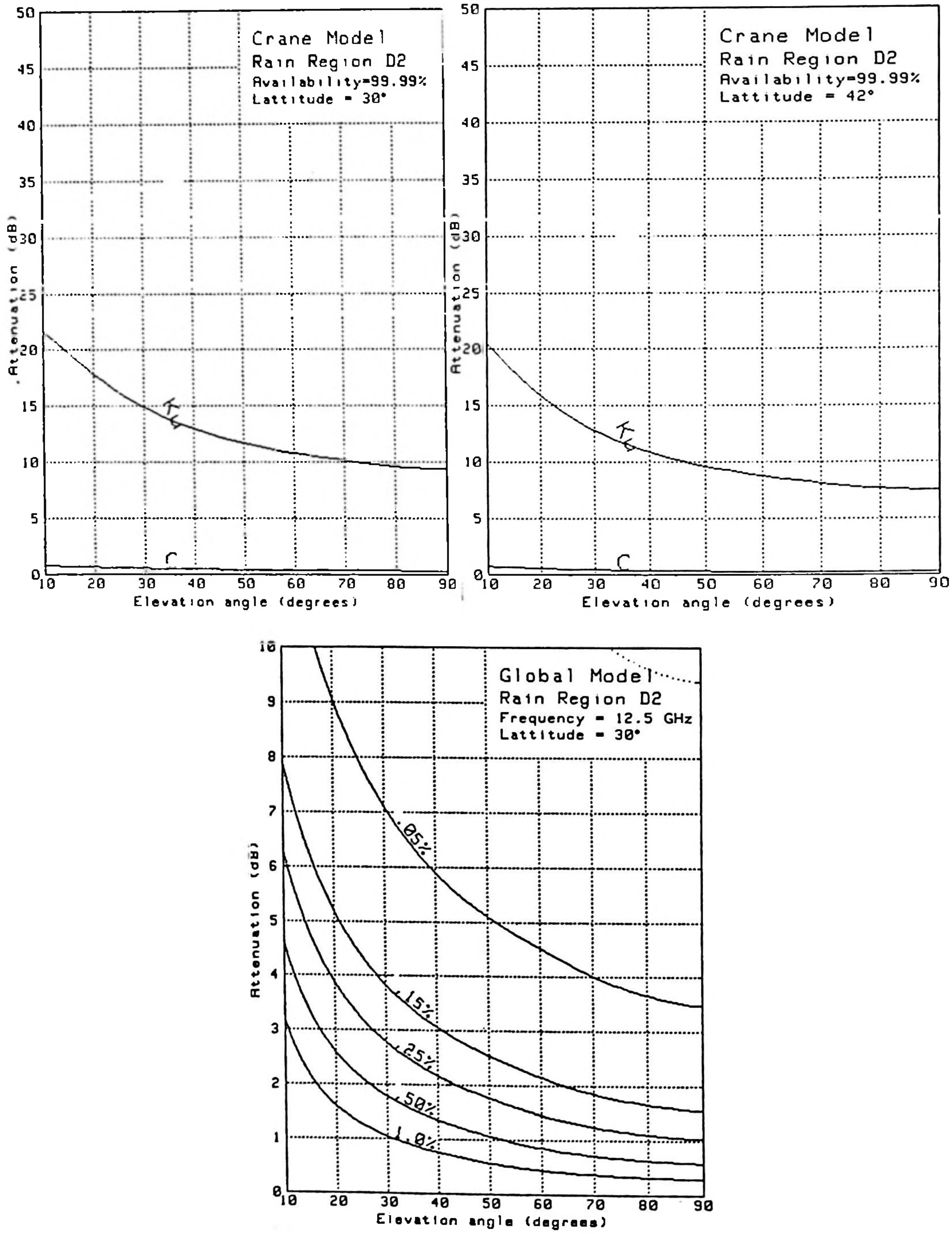
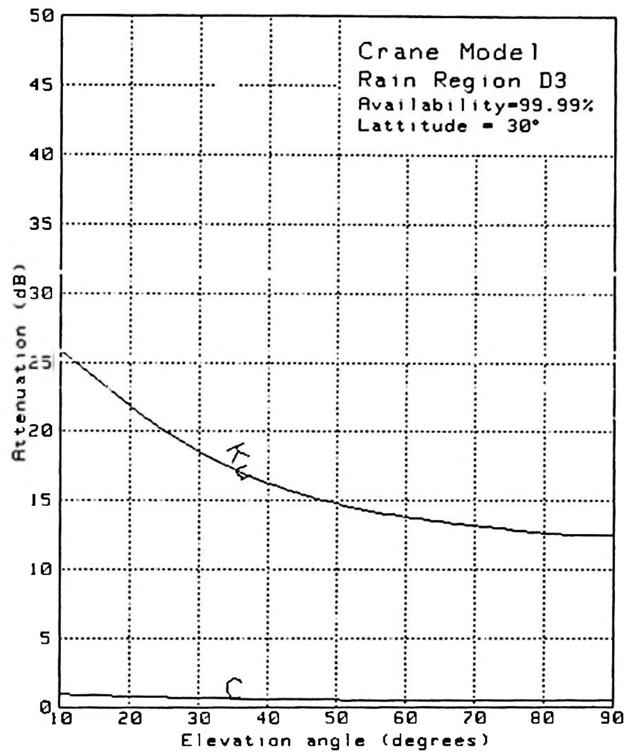


FIGURE 7. RELIABILITY DATA -- RAIN RATE REGION D2



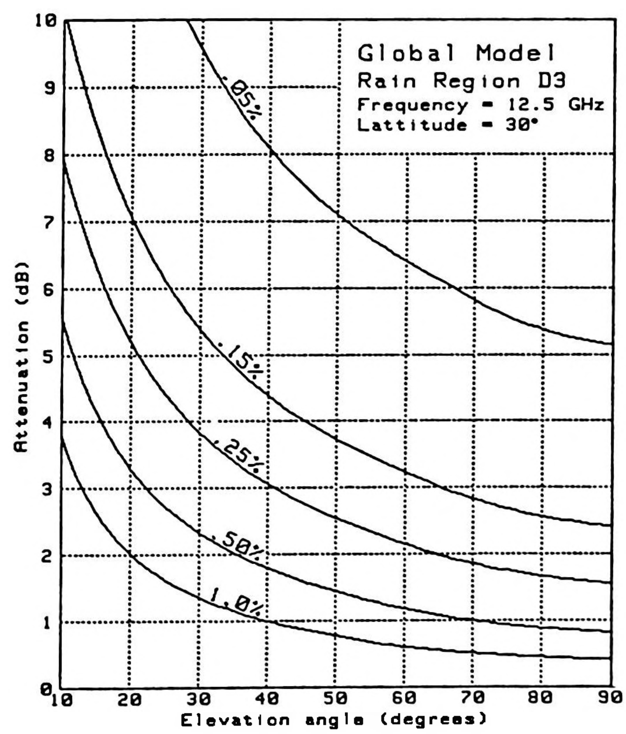


FIGURE 8. RELIABILITY DATA – RAIN RATE REGION D3

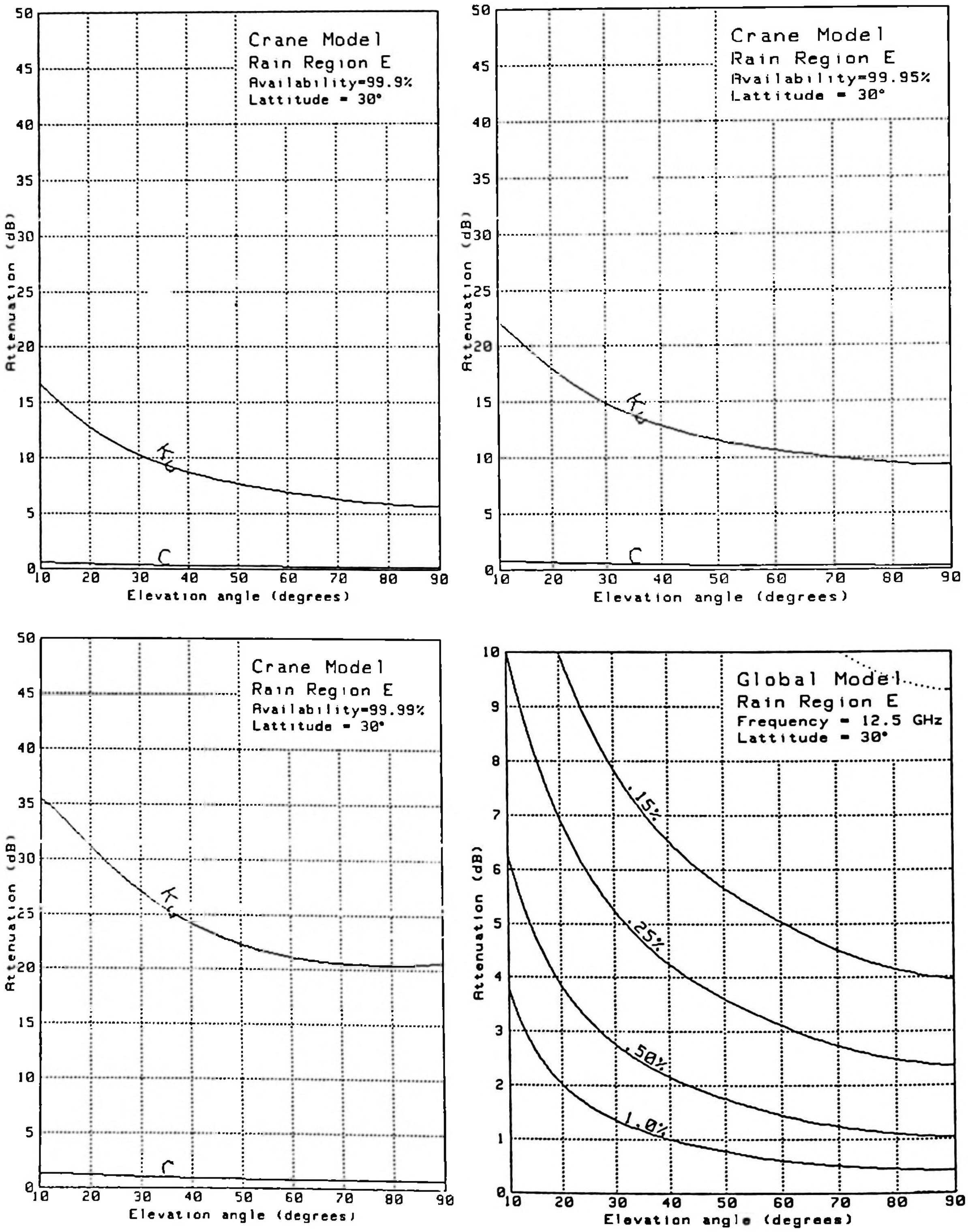


FIGURE 8. RELIABILITY DATA – RAIN RATE REGION E

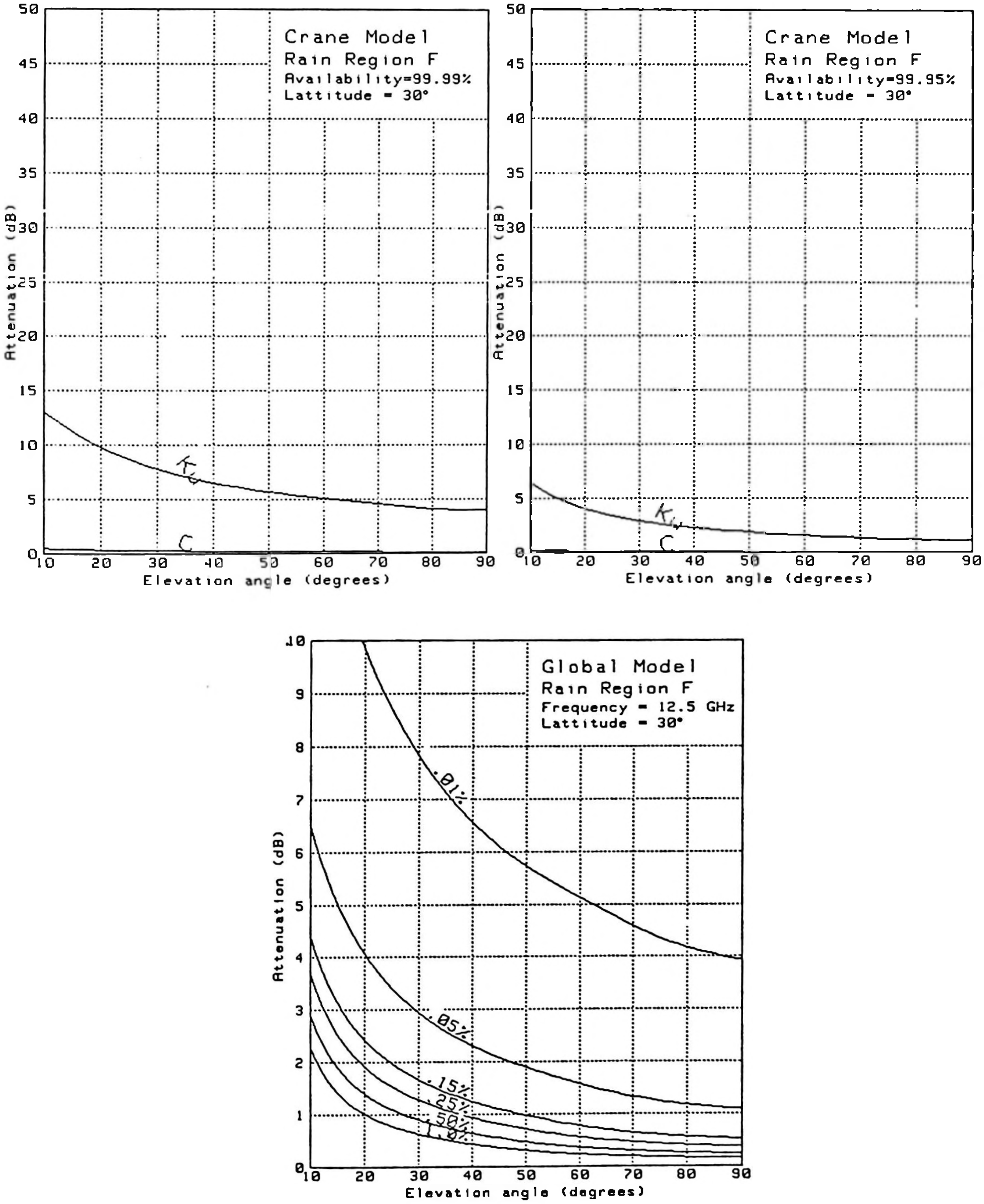


FIGURE 10. RELIABILITY DATA – RAIN RATE REGION F

TERRESTRIAL INTERFERENCE

C band satellite transmissions share the 6 and 4 GHz frequencies with common carrier microwave. This fact has not hindered development of many satellite delivery television networks. There are approximately 6000 cable TV antennas in place at C band. Public Broadcasting (TV and radio) has approximately 400 antennas in place. Experience with these networks indicates that about 90 percent of the cable antennas could be placed at their head ends. Of the Public Broadcasting antennas 75 percent of the installations for radio and 66 percent of the TV installations are colocated with their transmitters. The balance of the affiliates are served by including a terrestrial link.

In the paragraphs which follow, a plan is presented to modify the satellite link video signal structure and radio frequency bandwidths which effectively eliminates terrestrial interference as a consideration in the comparison of C or Ku band satellite transmission.

Both satellite and terrestrial systems have a fixed frequency plan. These plans are shown in Figure 11. It should be noted that interfering carriers from terrestrial systems into satellite channels will always show up ±10 MHz from the satellite carrier. These terrestrial carriers are narrowband and since they are frequency division multiplex (FDM)/FM carriers, the sideband distribution drops off rapidly at the edges of the band. It has been proven that a satellite TV transmission with 5 to 6 MHz deviation with an occupied bandwidth of 15 MHz (and suitable receivers) can operate in accordance with broadcast quality standards in an environment where the interfering carriers are stronger than the desired satellite carrier at the input to the receiver. This condition is shown graphically in Figure 11c.

This reduction in deviation and bandwidth results in lower signal-to-noise ratio for a given set of conditions. This shortfall can be readily made up with very little added expense. Consider the waveforms of Figure 12. The synch pulses in television transmission are very wasteful of satellite power. Thirty percent of the peak-to-peak voltage is in the synch. If the deviation and bandwidth were reduced in half, a 3 dB shortfall in signal-to-noise ratio would occur. This 3 dB can be made up by the simple expediant of the inversion of the synch pulse. The synch is easily inverted at both ends of the link.

Certain other advantages can accrue by either inverting the synch or substituting other data for the synch pulse. For example, the inverted synch by itself represents a mild form of scrambling. In addition, TV sound (audio) can be transmitted on the horizontal synch. Sound-on-synch has an additional benefit from the transmission standpoint. Elimination of the subcarrier for sound allows for wider deviation of the video in a given bandwidth, thereby improving signal-to-noise ratio.

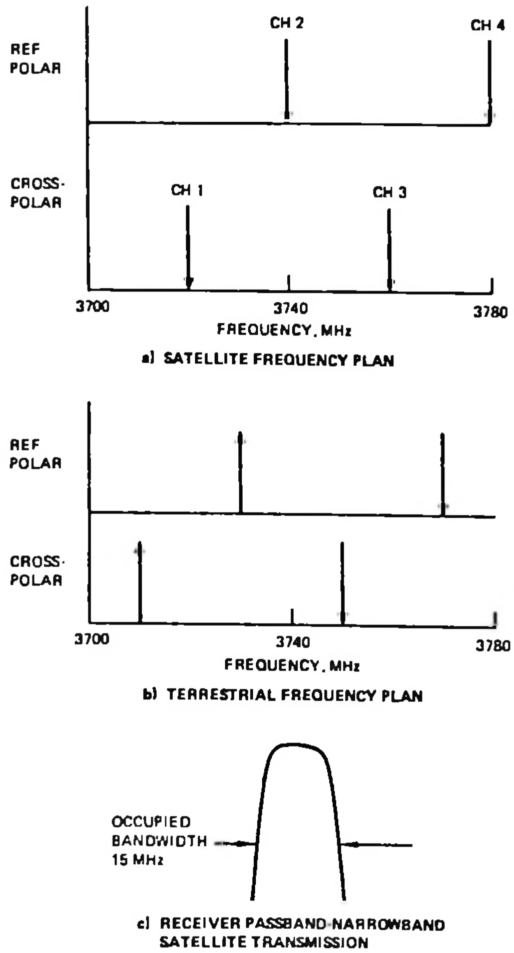


FIGURE 11. SATELLITE AND TERRESTRIAL FREQUENCY PLANS

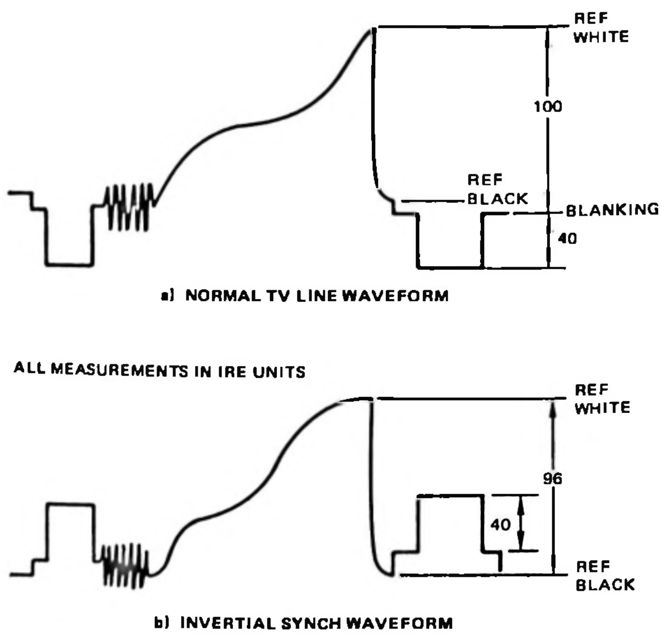


FIGURE 12. NORMAL AND INVERTED SYNCH WAVEFORMS

In addition, there are several systems currently available at a modest price which utilize the horizontal synch interval for a secure scrambling function as well as multichannel sound. These systems use less satellite power than a single subcarrier providing one sound channel.

CONCLUSIONS

In comparing C band and Ku band for U.S. domestic satellite delivery of "broadcast quality" television, C band is clearly more desirable. The advantage of C band is that it is virtually weather independent, allowing full U.S. coverage exceeding the 99.99 percent continuity of service objective. To meet the continuity of service objective at Ku band would be extremely expensive in a very large portion of the U.S. and prohibitive in certain regions of the U.S. Terrestrial interference into C band video transmission systems can be overcome by a judicious selection of modulation parameters, receiver bandwidth changes, and signal structure modification, which, with light additional complexity, can produce some desirable features such as security and multichannel sound.

This conclusion in no way deprecates the use of Ku band for other purposes. Other TV quality transmissions, digital TDMA, and narrowband data service for business may benefit from not having to share spectrum with terrestrial services. However, all applications should take into account the greater dependence of Ku band reliability on weather.

BIBLIOGRAPHY

1. R. Kaul, et al., A Propagation Effects Handbook for Satellite Systems Design, NASA Report N80-25520 (ORI, Inc. Technical Report 1679), March 1980.

2. R. K. Crane, "Prediction of Attenuation by Rain," IEEE Transactions on Communications, Vol. COM-28, No. 9, September 1980, pp. 1717-1733.

3. L. J. Ippolito, "Radio Propagation for Space Communications Systems," Proceedings of the IEEE, Vol. 69, No. 6, June 1981, pp. 697-727.

AN ANALYSIS OF INTERFERENCE IN SATELLITE TRANSMISSION OF VIDEO SIGNALS

A.G. Uyttendaele

AMERICAN BROADCASTING COMPANIES, INC.

New York, N.Y.

Abstract:

As is the case for all Radiocommunication services, satellite transmission performance is limited by the level of the noise which is superimposed on the received signal, as well as by the level of interference caused to the signal.

In order that the quality of the video signal be acceptable, the specification of the receive earth station system must therefore, include not only a minimum figure for the signal-to-noise, but also a minimum value for the interference ratio, which is the ratio, measured at the receiver input, of the power of the wanted signal to the sum of the powers of the interfering signals or C/I total.

In the long term, the interference problem is the more critical one, and it is essential that the receive earth station system be designed in such a manner as to minimize interference.

1. INTRODUCTION

The most important measure of picture quality in a satellite television link is the signal-to-noise ratio (S/N), defined as the ratio expressed in dB, of the peak-to-peak amplitude of the picture luminance signal to the RMS weighted noise. RS-250B, the EIA "Electrical Performance Standards for Television Relay Facilities," makes the following recommendations for signal-to-noise performance:

> "For a satellite circuit, including the uplink earth station transmitter - satellite transponder - downlink earth station receiver: The signal-to-noise ratio shall not be less than 56 dB. This compares to a S/N design objective of 54 dB for a "long-haul" terrestrial microwave relay system of up to 150 repeaters."

S/N is directly related to receive signal strength, and for every dB increase in receive signal level, the S/N will also improve by 1 dB. Receive signal strength, in turn, is directly related to antenna surface area and, therefore, when going from a 7 meter antenna to a 10 meter antenna, the antenna area is roughly doubled, resulting in a S/N improvement of about 3 dB.

The signal level available from a geostationary satellite, unlike that from terrestrial microwave sources, is power limited and extremely weak, but quite predictable. This is because the transmitter power and antenna gain of the satellites are well-defined and so is the path loss.

Video transponders aboard the current generation of DOMSATS have a relatively high EIRP - in the order of +63 dBm (33 dBw). Given a free-space path loss for 4 GHz in the order of -196 dB, the signal level available to the receive antenna will be about -133dBm

For a number of reasons, additional losses, such as from antenna mispointing errors, polarization misalignment, rainfall and outside interference influence the operation of every receive earth station. Good receive station performance can only be obtained if the system is designed with sufficient margin to account for these losses and interferences (noise).

2. INTERFERENCE ANALYSIS

This analysis is intended to demonstrate the susceptibility of both the satellite and receiving earth station to interference from many different sources such as:

1 - Internal interference of the satellite, primarily due to the two adjacent 20 MHz off-set cross polarized channels.
2 - Interference from the co-frequency channels of the adjacent satellites.
3 - Interference from other transmitting earth stations.
4 - Interference from terrestrial microwave links sharing the same frequency band.

Close attention should, therefore, be given to the question of what constitutes an acceptable level of interference, to the sources and type of interference, and on how to minimize these interferences.

It will become evident that the receive antenna performance characteristics of the earth station may be the only source of isolation between transponders on the same satellite and those on another satellite. The antenna selection, therefore, becomes an important decision in that it offers significant cost-performance trade-offs consistent with system performance requirements.

2.1. EARTH-STATION ANTENNA SIDELOBE REGULATION

An ideal antenna would radiate all its energy in a pencil beam with none escaping in any other direction. In practical antenna systems, most of the energy is in the main-beam, but significant levels of energy are radiated (or received) at angles off the main beam.

CCIR Report 391-1 established a formula to be used as a guide for peak sidelobe levels that might be expected in antennas used for earth stations. This formula, later incorporated in FCC Rules and Regulations §25.209 Antenna Performance Standards, states:

"Any antenna to be employed in transmission at an earth station in the Communication Satellite Service shall conform to the following standards:

Outside the main-beam, the gain of the antenna shall lie below the envelope defined by:

$$32-25 \log \theta \text{ dBi} \qquad 1^{\circ} \leqslant \theta \leqslant 48^{\circ}$$
$$-10 \text{ dBi} \qquad 48^{\circ} < \theta \leqslant 180^{\circ}$$

Where θ is the angle in degrees from the axis of the main lobe, and dBi refers to dB relative to an isotropic radiator.

For the purposes of this section, the peak gain of an individual sidelobe may be reduced by averaging its peak level with the peaks of the nearest sidelobes on either side, or with the peaks of two nearest sidelobes on either side, provided that the level of no individual sidelobe exceeds the gain envelope given above by more than 6 dB."

Using this 32-25 log θ formula, a typical pattern envelope for a 5 meter antenna is as shown in Figure 1. The on-axis gain for F = 3.95 GHz is about 44.5 dBi.

The gain for $\theta = \pm 1^{\circ}$ off-axis is 32-25 log 1 = 32 dBi or 12.5 dB below the gain at the peak of the beam. For $\theta = \pm 10^{\circ}$ off-axis, the gain is 32-25 log 10 = 7 dBi or 37.5 dB below the gain at the peak of the beam. This is graphically shown in Figure 2 for different antenna sizes, except that the horizontal axis is not labeled "Angle Off-Axis" but rather "Satellite-Spacing-Angle." This is justified since 4° of longitudinal separation at the geostationary orbit is essentially 4° as measured from an earth station antenna, and θ and the satellite separation angle are approximately the same. This is illustrated in Figure 3, where the geocentric spacing angle is shown as being equal to the topocentric spacing angle. In reality, the topocentric angle is always slightly greater than the geocentric spacing angle and therefore, calculations based on geocentric spacing angles are somewhat conservative.

In earth stations, the desired signal source is on the main beam axis, therefore, the curves of Figure 2 may be used to read the carrier signal to adjacent satellite interference ratio (commonly known as C/I), assuming that all satellites radiate the same EIRP in all directions, use identical modulation schemes, and, in effect, are homogeneous. The need to discriminate against unwanted signals will increase due to the growth in the number of satellites at smaller orbital spacings. Small earth station antennas have exacerbated the problem because such antennas have less off-axis discrimination, and, therefore, are far less able to suppress unwanted signals.

Interference from Adjacent Satellites occurs in two ways:

1 - Uplink interference from earth stations transmitting to adjacent satellites and,
2 - Downlink interference from adjacent satellites into the desired earth station.

The interference in both the uplink and downlink consists of several sources but is primarily caused by the co-frequency channels of the adjacent satellites. Figure 3 shows a hypothetical situation with three adjacent satellite systems. The spacing shown is 4^{o} and 5 meter antennas are used for transmit and receive purposes, meeting the present CCIR recommended sidelobe performance specifications.

Consider System #2:

Satellite #2 receiver picks-up interference from System #1 and System #3 earth station uplinks. The retransmitted signal from Satellite #2 contains both the wanted and interfering signals. Because of the earth station sidelobes of System #2, interference from Satellites #1 and #3 is picked-up. This is in addition to the wanted signal from Satellite #2 which is already contaminated. From the curves given in Figures 2 and 4, the following levels of interference can be determined:

TABLE #1
C/I dB

LINK	FREQ.	DUE TO SYSTEM 1	DUE TO SYSTEM 3	TOTAL C/I
Uplink	6 GHz	30.5	30.5	27.5
Downlink	4 GHz	27.7	27.7	24.7
System	-	-	-	22.8

TABLE #2
C/I in dB

LINK	FREQ.	DUE TO SYSTEM 1	DUE TO SYSTEM 3	TOTAL C/I
Uplink	6 GHz	29.3	29.3	26.3
Downlink	4 GHz	26.4	26.4	23.4
System	-	-	-	21.6

In the uplink, the C/I at the output of Satellite #2 receiver is about 30.5 dB due to one interfering earth station network, it degrades another 3 dB to 27.5 dB when two networks are involved, as can be read from Figure #2.

Due to the lower frequency, downlink discrimination is somewhat poorer than uplink discrimination and the on-axis signal is also reduced because of the lower gain of the receive antenna.

The C/I from either satellite is about 27.7 dB or 24.7 dB for both, as can be read from Figure #2. Combining the uplink and downlink interferences on a power basis yields an overall C/I of 22.8 dB and this will reduce the effective S/N ratio.

As satellites crowd the orbit and spacing between satellites may have to be reduced, earth stations may require to use more discriminating antennas, which means, better antennas (not necessarily larger antennas) will be required.

If we repeat this example but now use 10m antennas and space the satellites closer together, for example, to 2°, the following C/I ratios as given in Table #2 can be expected.

This example indicates that using 10 meter antennas with 2° spacing between satellites would result in a somewhat worse (1.2dB) C/I than using 5 meter antennas with the satellites spaced at 4°.

To improve the C/I when reducing the spacing between satellites, it will be necessary to improve the sidelobe performance of the antennas. In its Rulemaking Proceeding CC 81-704 on 2° spacing, the FCC recommends that the antenna pattern envelope be improved by 3 dB at small off-axis angles between 1° and 7°. This improvement would change the numbers in Table #2 to those of Table #3:

LINK	C/I in dB			Total C/I
	FREQ.	DUE TO SYSTEM 1	DUE TO SYSTEM 3	dB
Uplink	6 GHz	32.3	32.3	29.3
Downlink	4 GHz	29.4	29.4	26.4
System	-	-	-	24.6

Or as could be expected, a 3 dB improvement in the C/I over the previous example.

Figures 5 & 6 show the present and proposed antenna pattern envelopes for 4 GHz and 6 GHz for a 10 meter antenna.

Present

$32-25 \log \theta$ dBi $\quad 1^{\circ} \leqslant \theta \leqslant 48^{\circ}$
-10 dBi $\quad 48^{\circ} < \theta \leqslant 180^{\circ}$

Proposed

$29-25 \log \theta$ dBi $\quad 1^{\circ} \leqslant \theta \leqslant 7^{\circ}$
$+8$ dBi $\quad 7^{\circ} < \theta \leqslant 9.2^{\circ}$
$32-25 \log \theta$ dBi $\quad 9.2^{\circ} < \theta \leqslant 48^{\circ}$
-10 dBi $\quad 48^{\circ} < \theta \leqslant 180^{\circ}$

As can be seen from the 6 GHz curves, for 4^{o} spacing, the sidelobe level is 37 dB down from the peak of the beam for the present antenna requirements. With the proposed pattern envelope, this would improve to 40 dB. It should, therefore, be possible to space the satellites closer together and as can be seen from the graphs, the status quo would be maintained for 3^{o} spacing.

Why then would the FCC propose a 2^{o} spacing when the antenna pattern envelope improvement of 3 dB as proposed would only allow the same interference levels at 3^{o} as are presently available at 4^{o}? To understand this, the FCC also proposes to improve the polarization isolation and to use a satellite polarization interleaving plan of co-frequency transponders to pick-up the additional isolation required, as explained below. Again, from the graph in Figure 6, it is apparent that to obtain the same 37 dB C/I at 2^{o} spacing as at 4^{o} spacing, the pattern envelope has to be improved an additional 4.5 dB from the already 3 dB proposed.

It would, therefore, be possible to retain the present C/I of 4^{o} for 2^{o} spacing if the pattern could be improved from 32-25 log θ to 24.5 - 25 log θ.

2.2. POLARIZATION DIVERSITY/FREQUENCY REUSE

The frequency spectrum available for satellite transmission in the 4/6 GHz band is 500 MHz wide, from 3,700 to 4,200 MHz for the downlink and from 5,925 to 6,425 MHz for the uplink. This 500 MHz wide spectrum is divided into 12 individual channels each 40 MHz wide. A typical frequency plan is as shown in Figure 7 and this is the frequency plan used by the first WESTAR satellites.

To effectively double the channel capacity while still using only the 500 MHz bandwidth available at 4 and 6 GHz, the newer satellites use polarization diversity as a means of reusing the frequency spectrum.

A polarization diversity technique could consist of transmitting 12 signals in one polarization (vertical) and 12 signals to an orthogonal polarization (horizontal) to and from the satellite.

To further improve the polarization isolation between signals, the horizontal and vertical polarized channels are interleaved with each other as shown in Figures 8, 9, 10 for the SATCOM, COMSTAR and WESTAR IV & V satellites. The horizontal and vertical channels are off-set from each other by 20 MHz.

All of the 36 MHz wide channels are spaced 40 MHz apart resulting in a 4 MHz guardband between adjacent channels. These guardbands complement the polarization isolation inherent in the antenna design.

Because of less than ideal antennas and transmission media, this type of a system is especially susceptable to interference from the interleaved cross-polarized signals since a portion of these signals occupy the same frequency band as the desired signal as shown in Figure 11. To minimize the interference from cross-polarized signals, it is important to achieve high discrimination between the polarizations at both the transmitting and receiving ground stations as well as in the satellite.

For the ground station antenna, it is essential to maintain the polarization purity at the peak of the beam. The spacecraft must, however, maintain polarization purity over the entire shaped beam or over a 3 degree x 7 degree contour for CONUS.

To maintain the required isolation between channels, an overall system polarization isolation of about 25 dB needs to be maintained. To achieve this, the budget for the AT&T system requires 35 dB polarization isolation at the earth stations and 33 dB at the satellite.

Cross polarization may, however, be relaxed dB for dB by the amount that the transmitted carrier is less than required for transponder saturation. To access the SATCOM satellites, the earth station isolation between orthogonal cross-polarized signals shall be at least 39 dB throughout the band of 5.925 to 6.425 GHz.

2.3 FARADAY ROTATION

The most devastating transmission media phenomena that impacts polarization isolation are rain and Faraday rotation.

The orientation of linearly polarized signals passing through the ionosphere (altitude of 50 to 400 km) is affected by the slowly varying magnetic fields that encompass the earth: Faraday rotation. These slowly varying fields must be tracked and compensated for.

Faraday rotation is not a deorthogonalizing effect, this means the orthogonality of dual linearily polarized signals is not impacted by Faraday rotation. The polarization merely rotates and may, therefore, be compensated for by rotation of the transmitting or receiving antenna feed.

In the northern hemisphere, the uplink rotation is CCW and the downlink rotation is CW with respect to the direction of propagation. The peak rotation at noon can be approximated by $R = 90/F^2$ (degrees) where F is the signal frequency in GHz. The minimum rotation at dawn is approximately $R = 18/F^2$.

Interference due to angular misalignment is equal to 20 log sin Ø, where Ø is the rotation angle. For a misalignment of only one degree, isolation drops to 36 dB and if the orthogonal signals are off by 4.6 degrees, only 22 dB of isolation is theoretically available.

2.4. ADJACENT SATELLITE INTERFERENCE FOR FREQUENCY REUSE MODEL

As shown in Figure 3, interference from adjacent satellites occurs in two ways:

1 - Uplink interference from earth stations transmitting to adjacent satellites and,
2 - Downlink interference from adjacent satellite transmission into the desired earth station.

For the frequency reuse model of Figure 13, the interference in both the uplink and downlink consists of many signals but is primarily caused by the co-frequency channels and the two 20 MHz off-set frequency channels.

Consider System #2:

The signals from all odd-channels for System 2 are horizontally polarized whereas the even-channels are vertically polarized. The odd-channels for System 1 and 3 are vertically polarized and the even-channels are horizontally polarized.

The polarization discrimination of the earth station antennas at the peak of the beam is at least 35 dB. If now the polarization discrimination remains excellent for off-axis angles, then the interference from the two co-frequency signals from the adjacent satellites received by #2 satellite will be further reduced. However, the overlapping channels F_2 & F_4 of System #1 and #3 are co-polarized with the signal F_3 of System #2 and no further discrimination is achieved.

This is the frequency/polarization interleaving satellite model proposed by the FCC in its Rulemaking Proceeding CC 81-704 on 2° spacing. The new antenna performance standards would be: (§25.209).

(a) The gain of any antenna to be employed in transmission from an earth station in the fixed-satellite service shall lie below the envelope defined by:

$29-25 \log \theta$ dBi	$1^{\circ} \leqslant \theta \leqslant 7^{\circ}$
$+8$ dBi	$7^{\circ} < \theta \leqslant 9.2^{\circ}$
$32-25 \log \theta$ dBi	$9.2^{\circ} < \theta \leqslant 48^{\circ}$
-10 dBi	$48^{\circ} < \theta \leqslant 180^{\circ}$

where θ is the angle in degrees from the axis of the main lobe, and dBi refers to dB relative to an isotropic radiator. For the purposes of this section, the peak gain of an individual sidelobe may not exceed the envelope defined above by more than 3 dB for θ between 1° and 7°. For θ

greater than 7°, the envelope may be exceeded by 10% of the sidelobe peaks.

(b) The off-axis cross-polarization isolation of any antenna to be employed in transmission from an earth station in the fixed satellite service shall be greater than or equal to 10 dB for off-axis angles between 1° and 10°.

Figure 14 and 15 show graphs for the present and proposed performance of a 10 meter antenna for 4 GHz and 6 GHz.

Curve #1	$32-25 \log \theta$	pattern envelope
Curve #2	$26-25 \log \theta$	estimated X-pol isolation
Curve #3	$29-25 \log \theta$	proposed pattern envelope
Curve #4	$19-25 \log \theta$	proposed X-pol isolation

Curve #4 is the response to cross-polarized signals from adjacent satellites. The difference between Curve #3 and #4 is called: The off-axis cross-polarization discrimination.

Present antennas are estimated to have 4 to 6 dB off-axis cross-polarization discrimination. This would have to be improved to 10 dB for off-axis angles between 1° and 10° as proposed by the FCC in its Rulemaking Proceeding on 2° spacing.

Figure 14 shows that the interference signal level from an adjacent satellite at 4° is 40 dB below the level of the desired signal using the present performance requirements. It is also 40 dB for 2° spacing if the antenna meets the proposed sidelobe performance and the proposed improved cross-polarization isolation of 10 dB.

2.5. CONCLUSION

In order to achieve closer spacing of satellites, would require for satellites having orthogonal polarization and frequency interleaved transponders on uplinks and downlinks to be located adjacent to one another, i.e., transponder 7 on one satellite would be vertical on the uplink and horizontal on the downlink, whereas in the adjacent satellites, transponder 7 would be horizontal on the uplink and vertical on the downlink. As can be seen from Table 4, the SATCOM IIIR and WESTAR IV & V are so designed. In addition, the satellites have to be homogeneous, i.e., the interfering and desired satellites should have the same saturation flux density and radiated EIRP. (The radiation pattern should yield the same signal strength at any given location on the ground for adjacent satellites).

TABLE #4

	COMSTAR*	SATCOM	WESTAR IV & V
Polarization, Downlink			
3720,3760,3800, etc.	V	V	H
3740,3780,3820, etc.	H	H	V
Polarization, Uplink			
5945,5985,6025, etc.	V	H	V
5965,6005,6045, etc.	H	V	H

* The TELSTAR 3 satellites, presently under construction, are to be compatible with the COMSTAR satellites which they will replace and will employ the COMSTAR transponder frequency and polarization plan. This permits direct replacement of a COMSTAR satellite with a TELSTAR 3 series satellite without change to the system's earth stations or to the operation of the system.

It would not be possible to realize these benefits in a systematic way unless and until preferred polarization characteristics have been adopted. Implementing an improved satellite plan will, therefore, take several years. As presently, the three common carriers use three different frequency and polarization plans.

3. PRESENT DAY ENVIRONMENT

ABC's initial transponder is 7V on COMSTAR D3 at 87°W longitude. Adjacent to COMSTAR D3 at 83° is RCA's SATCOM IV and at 91°, Western Union's WESTAR III. See Figure 16.

The current frequency/polarization model for all transponders, co-channel, 20 MHz off-set and 40 MHz off-set is as shown in Figure 17 and Table 5.

	SATCOM IV(83°)	COMSTAR D3(87°)	WESTAR III(91°)
FREQUENCY	UPLINK	UPLINK	UPLINK
6145 GHz	11H	6V	6V
6165 GHz	12V	6H	-
6185 GHz	13H	7V	7V
6205 GHz	14V	7H	-
6225 GHz	15H	8V	8V
FREQUENCY	DOWNLINK	DOWNLINK	DOWNLINK
3920 GHz	11V	6V	6H
3940 GHz	12H	6H	-
3960 GHz	13V	7V	7H
3980 GHz	14H	7H	-
4000 GHz	15V	8V	8H

SATCOM IV and COMSTAR D3 employ orthogonal polarization and thus are frequency reuse satellites, each having 24 transponders whereas WESTAR III is a 12 transponder satellite.

An interference model can now be drawn which depicts the interference contribution for the uplink and downlink and the overall system C/I can then be calculated for different receive antenna sizes for different locations in the country, taking into account the different EIRP's.

The following assumptions, based on published data, are made:

1 - All uplink antennas are 10 meters and meet the 32-25 log θ sidelobe requirements and a polarization isolation of 6 dB off-axis is assumed.

2 - All transponders are saturated at -82 dBw/m^2 power flux density for SATCOM IV and WESTAR III, and at -76 dBw/m^2 PFD for the COMSTAR D3 transponders.

3 - Because of the partially overlapping frequency spectra of the two 20 MHz off-set-frequency transponders, a frequency improvement factor "Fi" relating to the spectra of the desired and undesired signals has to be considered. If both transponders carry FM/TV signals occupying the same bandwidth, this frequency discriminating factor would be 0 dB. For a 20 MHz off-set-frequency FM/TV signal, interfering with another FM/TV signal, this term is usually taken as Fi = 6.5 dB.

<u>Interference into System #2 of Figure 17</u>

Interference into the wanted system usually originates from three major sources:

1 - Adjacent satellite interference. (C/I_{adj})
2 - Internal interference within the satellite. (C/I_{int})
3 - Terrestrial microwave interference. (C/I_{terr})

Total carrier-to-interference C/I_{tot} defined as the ratio, in dB, between the power of the wanted signal and the power of all interfering signals as measured at the input of the earth station receiver is expressed in equation form as:

$$C/I_{tot} = C/I_{adj} \oplus C/I_{int} \oplus C/I_{terr}$$

<u>Adjacent Satellite Interference</u>

Interference from adjacent satellites may occur in two ways:

- Uplink interference from earth stations transmitting to adjacent satellites and,

- Downlink interference from adjacent satellites transmitting into the desired earth station.

The primary interferers are:

- The co-frequency channels from the first adjacent satellite on each side. These channels can be co-polarized or cross-polarized.
- The two 20 MHz off-set-frequency channels from the first adjacent satellite on each side. These channels can be co-polarized or cross-polarized.

If a more accurate model is required, the contributions from these same channels for the second adjacent satellite would have to be included.

<u>Uplink Carrier-to-Interference Ratio:</u>

$$C/I_u = SFD_u \text{ (on-axis,wanted)} - SFD_u \text{ (off-axis,unwanted)}$$
$$= SFD_u \text{ (wanted)} - [SFD_u \text{ (on-axis,unwanted)} - \Delta G - Pi - Fi]$$

Where SFD_u = Saturated Flux Density, uplink in dBw/m^2

ΔG = off-axis gain differential of antenna in dB which for 4^o is equal to 37 dB and for 8^o equal to 44 dB.

Pi = Polarization Improvement, which for like polarization is 0 dB and for cross-polarization is 6 dB.

Fi = Frequency off-set, F, improvement factor, which for ΔF = 0 MHz, is 0 dB
for ΔF =20 MHz, is 6.5 dB

<u>C/I_u from Uplink for Satellite at 83^o</u>

1 - Co-frequency transponder, cross polarized:
$-76 - [-82 -37 -6] = 49$ dB

2 - Off-set frequency transponder, co-polarized:
(for each transponder)
$-76 - [-82 -37 -6.5] = 49.5$ dB

<u>C/I_u from Uplink for Satellite at 91^o</u>

Co-frequency transponder: $-76 - [-82 -37] = 43$ dB

Total C/I_u is: $49 \oplus 49.5 \oplus 43 = \underline{40.7}$ dB

This C/I is excellent and is realized because of the high SFD required to saturate the D3 transponder. If the SFD were reduced to -82 dBw/m^2, the overall C/I_u would also reduce by 6 dB or to <u>34.7</u> dB.

<u>Downlink Carrier-to-Interference Ratio:</u>

$$C/I_d = \left[EIRP_d + G_R\right] \text{(wanted signal)} - \underbrace{\left[EIRP_d - G_R(\theta_i) - F_i - P_i\right]}_{\text{(unwanted)}}$$

Where: G_R: Gain of receive antenna.

$G_R(\theta_i)$: Gain of receive antenna at off-axis angle=θ_i

F_i: Frequency-offset improvement factor equal to 6.5 dB for 20 MHz offset.

P_i: Cross-polarization improvement, equal to 6 dB for 4^o off-angle.

For equal EIRP's from all three satellites, the $\Delta G_R(\theta_i)$ can be read from the graphs of Figure 2.

$$C/I_d = \Delta G_R + F_i + P_i$$

	10m	7m	5m
1 - Co-frequency, co-polarized channel at 83^o.	34 dB	31 dB	28 dB
2 - 20 MHz offset-frequency, cross polarized channels at 83^o: 34 dB + 6.5 dB + 6 dB each for a 10 meter antenna.	43.5dB	40.5dB	37.5dB
3 - Co-frequency, cross polarized channel at 91^o.	40 dB	37 dB	34 dB
C/I_d =	32.6dB	29.6dB	26.6dB

For every dB the EIRP of the unwanted satellite is higher than that of the wanted, the C/I_d will degrade by 1 dB and conversely, for every dB the EIRP of the unwanted satellite is less than that of the wanted, the C/I_d will improve by 1 dB.

Satellite Internal Interference:

The internal interference of a frequency reuse satellite is primarily due to the two adjacent 20 MHz offset-frequency cross-polarized channels of the satellite.

Assuming that the uplink/downlink antennas in conjunction with the satellite antenna meet the 28 dB cross polarization performance objective, then the carrier-to-interference caused into Channel 7V when uplinking to transponder 6V is: 28 dB + 6.5 dB = 34.5 dB. The same C/I can be expected when uplinking to transponder 7H and, therefore, the combined C/I is 31.5 dB.

For the downlink, equal C/I values can be expected and, therefore, the combined C/I will be 28.5 dB. A C/I_{int} = 26 dB for frequency reuse satellites is considered typical.

System C/I:

$$C/I_{tot} = C/I_u \oplus C/I_d \oplus C/I_{int}$$

$= 40.7$ dB $\oplus$ 32.6 dB $\oplus$ 28.5 dB = 26.9 dB for 10m

$= 40.7$ dB $\oplus$ 29.6 dB $\oplus$ 28.5 dB = 25.9 dB for 7m

$= 40.7$ dB $\oplus$ 26.6 dB $\oplus$ 28.5 dB = 24.3 dB for 5m

Terrestrial Interference: C/I_{terr}

Terrestrial radio relay systems share a common frequency band between 3700 and 4200 MHz with the Fixed Satellite Service. The terrestrial microwave carriers are centered on frequencies offset by 10 MHz from the satellite carriers. C/I_{terr} = 25 dB is a typical level for which frequency coordination is usually accomplished.

For example, For C/I_{terr} = 25 dB, the new system C/I becomes; 40.7 dB $\oplus$ 32.6 dB $\oplus$ 28.5 dB $\oplus$ 25 dB = 22.8 dB for 10m, 22.4 dB for 7m, and 21.6 dB for a 5m antenna.

This analysis shows the importance of selecting a "quiet" site for the receive earth station, as the terrestrial interference may be the limiting factor in the overall performance.

4. PROTECTION RATIO

Protection ratio is the minimum value of the interference ratio $(C/I)_{min}$ that will give satisfactory reception.

Protection ratio can only be determined by subjective evaluation of picture quality in the presence of interference of known intensity. The effects of interference in TV transmissions are complex and are not easily analyzed.

Many experiments have been carried out by as many groups trying to relate the subjective measure of interference (viewing quality) to engineering terms such as carrier-to-interference ratio (C/I).

CCIR Report 634-1 provides an empirical expression showing the relationship between protection ratio PR, peak-to-peak frequency deviation D in MHz and quality impairment Q for two FM-TV signals of similar modulation characteristics assuming co-frequency operation.

$$PR = 13.5 - 20 \log (D/12) - Q + 1.1\, Q^2 \text{ dB}$$

Where D = 21.5 MHz or 27.6 MHz peak-to-peak deviation.
Q = Impairment grade, measured on a 5 point scale as described in CCIR Recommendation 500-1.

This scale is as follows:

TABLE #6

	Quality	Impairment
Q=5	Excellent	Imperceptible
Q=4	Good	Perceptible but not annoying
Q=3	Fair	Slightly annoying
Q=2	Poor	Annoying
Q=1	Bad	Very annoying

From the empirical expression for PR and Table #6, it is now possible to derive Table #7.

Q	C/I(dB) D=23.5 MHz	C/I(dB) D=27.6 MHz
5.0	30.9	28.8
4.5	26.2	24.0
4.0	22	19.9
3.5	18.4	16.2
3.0	15.3	13.2

In a 1971 report, NASA concluded that a protection ratio on the order of 27 dB is needed for high quality video transmission.

The FCC, on the other hand, in its NOI-CC 81-704 on 2^{o} spacing indicated that a C/I total of 18 dB is satisfactory for Cable Television as it provides acceptable TV reception and this value has been endorsed by the NCTA.

5. SATELLITE LINK ANALYSIS

The equation for downlink carrier-to-noise ratio C/N_d is:

$$C/N_d = EIRP - S + G/T - 10 \log B_{Hz} + 228.6 \qquad (1)$$

Where: EIRP is the Effective Isotropic Radiated Power, expressed in dBW (dB above one Watt), or satellite power transmitted toward a spot on the surface of the earth. "Footprint" maps show contours of constant EIRP.

S is Space Loss. A signal transmitted from the satellite to earth will experience considerable attenuation, on the order of 196.5 dB. The clear sky value of space loss is:

$$S = 123.54 + \log F_{MHz} + 10 \log \left[1.42 - 0.42 \cos (LAT) \times \cos (SAT\text{-}LONG)\right] \qquad (2)$$

Where: F_{MHz} is the carrier frequency in MHz. Between 3700 and 4200 MHz.

$\underline{LAT}$ is Receive Earth Station Latitude.
$\underline{LONG}$ is Receive Earth Station Longitude.

$\underline{G/T}$: Is referred to as the quality factor of the Antenna and LNA system.

$G/T = G_A - 10 \log T_s$ Where: $\underline{G_A}$ is the gain of the Antenna in dBi (3)

T_s is the total system noise temperature in ^oK.

$$T_s = T_A + (L_1-1).T_o + L_1.T_{LNA} \qquad (4)$$

Where: $\underline{T_A}$ is Antenna Noise Temperature in ^oK.

See Figures 18 & 19 for 4.5m and 10m antennas.

$\underline{T_o}$ = 290^{o}K on a typical day.

$\underline{T_{LNA}}$ = LNA Noise Temperature in ^oK.

L_1 = power loss ratio between antenna output port and LNA input, such as is caused by a waveguide switch in case of redundant LNA operation. Typically 0.1 dB.

Converted to power ratio: $L_1=10^{0.01}=1.0233$. Hence: $G/T = G_A - 10 \log T_A + (L_1-1).T_o + L_1.T_{LNA}$ (5)

Magnitude of bracketed terms:

T_A: See Figures 18 & 19 for 4.5m and 10m antennas.

$(L_1-1).T_o = 0.0233x290 = 6.755$

$L_1.T_{LNA} = 1.0233x80 = 81.863$ for an 80^{o}K LNA.

$= 1.0233x140 = 143.261$ for a 140^{o}K LNA.

T_s can, therefore, also be written as:

$T_s = T_A + T_{LNA} + T_{MISC} = T_A + 80 + 8.62$ for an 80^{o}K LNA

$= T_A + 140 + 10.02$ for a 140^{o}K LNA

(5) simplifies to: $G/T=G_A-10 \log\left[T_A+T_{LNA}+10\right]$ (6)

Figure 20 is a graphic representation of this expression. A desired value of G/T can be obtained by using a certain size antenna and LNA combination.

Examples:	For a 10m antenna and 80^{o}K LNA	G/T = 30.22
	For a 10m antenna and 140^{o}K LNA	G/T = 28.28
	For a 7m antenna and 80^{o}K LNA	G/T = 27.04
	For a 7m antenna and 140^{o}K LNA	G/T = 25.17

$\underline{B_{Hz}}$: Is the receiver pre-detection bandwidth in Hz, typically $32.8x10^6$ Hz for 10.75 MHz deviation and $36x10^6$ Hz for 13.8 MHz. Therefore: $10 \log B_{Hz}$ = 75.16 for 32.8 MHz BW
= 75.56 for 36 MHz BW

and (1) becomes: C/N_d = EIRP+G/T-43.06 for 32.8 MHz BW
C/N_d = EIRP+G/T-43.46 for 36 MHz BW

228.6: Is Boltzmann's constant, a conversion factor.

The signal delivered to the LNA/receiver for processing is not only determined by C/N_d but by $C/N_d \oplus C/I$.

Where C/I is the carrier-to-interference ratio.

Interference sources are: adjacent satellite signals being received by the receive earth station, interference generated internal to the satellite or 4 GHz terrestrial microwave signals.

$C/N_T = C/N_d \oplus C/I$

The video signal-to-noise ratio S/N_V for a carrier-to-noise signal level above receiver threshold is determined from C/N_T as follows: $S/N_V = C/N_T + K$ dB (7)

Where K is a constant which takes into account the following parameters, all basically non-controllable.

- Pre-detection Bandwidth — B = 32.8 MHz or 36.0 MHz
- Highest Video Frequency — f_V = 4.2 MHz
- Carrier Deviation — D = 10.75 MHz or 13.8 MHz
- Pre-De-emphasis Improvement — PDI
- Noise Weighting Factor — NW

PDI + NW: 12.8 dB

or:

$$K = 10 \log 6 + 10 \log (B/f_V) + 20 \log (D/f_V) + PDI + NW \quad (8)$$

K = 37.67 dB for 10.75 MHz deviation and 32.8 MHz BW
K = 39.84 dB for 13.80 MHz deviation and 32.8 MHz BW
K = 40.24 dB for 13.80 MHz deviation and 36 MHz BW

CONCLUSION

From equations (1) and (6), it becomes apparent that the only available method of controlling the receive signal quality is by controlling the system G/T. G/T is the figure of merit of the receive system. It is a function of the gain of the antenna, noise temperature of the LNA along with the antenna noise temperature and losses located ahead of the LNA. Improving G/T means increasing the gain of the antenna or decreasing the noise temperature of the LNA.

Example: The objective is to obtain a S/N_V equal to 56 dB; S/N_V is related to C/N by:

$S/N_V + C/N_T + 37.67$ from which $C/N_T = 56 - 37.67 = 18.33$ dB
$S/N_V + C/N_T + 39.84$ from which $C/N_T = 56 - 39.84 = 16.16$ dB
$S/N_V + C/N_T + 40.24$ from which $C/N_T = 56 - 40.24 = 15.76$ dB

The C/I levels are assumed to be respectively 22 dB, 25 dB and 28 dB and, therefore, C/N_d is as given in Table #8.

TABLE #8

	C/I=22dB	C/I=25dB	C/I=28dB	C/I= ∞
C/N_d	20.77	19.38	18.83	18.33
C/N_d	17.47	16.77	16.45	16.16
C/N_d	16.94	16.31	16.03	15.76

From (1) follows that:

$G/T = C/N_d$ - EIRP + 43.06 for 32.4 MHz BW
$G/T = C/N_d$ - EIRP + 43.46 for 36 MHz BW

TABLE #9

G/T for EIRP = 35 dBW and S/N_v = 56 dB

DEV.&BW	C/I=22dB	C/I=25dB	C/I=28dB	C/I= ∞
10.75/32.8	28.83	27.44	26.89	26.39
13.8/32.8	25.53	24.83	24.51	24.22
13.8/36	25.40	24.77	24.49	24.22

This is illustrated in Figure 21.
Figure 22 shows G/T for EIRP = 33 dBW and S/N_v = 56 dB while Figure 23 shows G/T for EIRP = 33 dBW and S/N_v = 54 dB.

USE OF NOMOGRAPHS

Suppose the Satellite EIRP = 35 dBW as determined from the satellite "footprint" and the objective is to obtain a S/N_v = 56dB. What size antenna and type LNA would be required?

All the dash-dot lines, "a" through "g" represent lines of constant EIRP = 35 dBW for different values of C/I and for different values of carrier deviation. These lines intersect the antenna/LNA curves as shown in Table #10.

DEV./BW	C/I = 22 dB	C/I = 25 dB	C/I = 28 dB
10.75/32.8	10m/120° 9.2m/100°	9.2m/152° 7m/70°	7m/83°
13.8/32.8	7m/125° 6.1m/93°	7m/153° 6.1m/114°	6.1m/126°

Therefore: For C/I = 22 dB and 25 dB, two choices are available of which the second will be the least expensive.

Acknowledgement: The author wishes to thank John Serafin for his valuable suggestions and assistance in editing this paper and Bonni Brodsky for her assistance and patience in preparing this manuscript.

References:

1. Richard T. Davis, "Satellite Communication: The Search for More Bandwidth," Microwaves (Jan.1975), pp. 14-18.
2. FCC NOI Docket #81-704 and comments by AT&T, Comsat, RCA and Scientific-Atlanta, (November 18, 1981).
3. M. Spellman, "Lower Crosspole Levels for Safe Frequency Reuse," Microwaves (Nov. 1977), pp. 48-58.
4. James H. Cook, "Expand System Capacity with Frequency Reuse," Microwaves (April 1978), pp. 94-96.
5. Walter L. Morgan, "Ever Smaller Earth Stations, Crowded Satellites Imply Changes," MSN (April 1979), pp. 81-90.
6. H. Paul Shuch, "A "Vidiots" Guide to Microwave TV Links," Microwaves (June 1979), pp. 40-46.
7. C. Louis Cuccia, "Regulations Direct Satcom Antennas," MSN (March 1980), pp. 45-57.
8. C. Louis Cuccia, "Satellite Communications and the Information Decade: An Overview," Microwave Journal, (Jan. 1982), pp. 22-34.
9. W. Braun, "Reducing Satellite Spacing," Satellite Circuit #1 (1982), pp. 15-17.
10. "Scientific Atlanta", Satellite Communications Symposium (1982).
11. "TVRO Satelllite Earth Station Handbook," Harris Corporation.

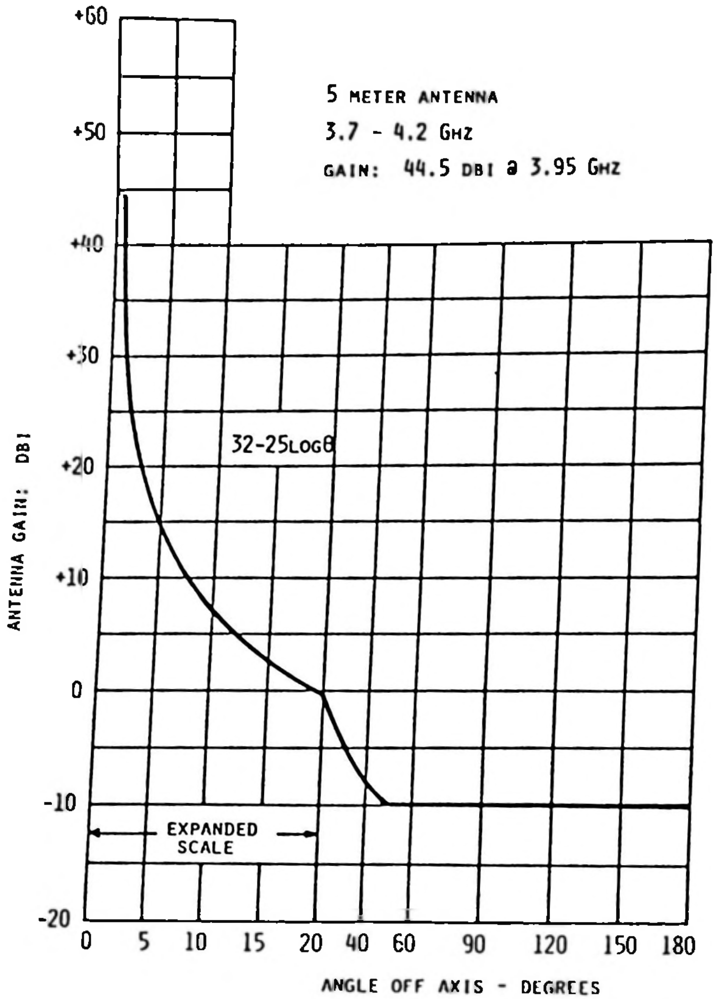
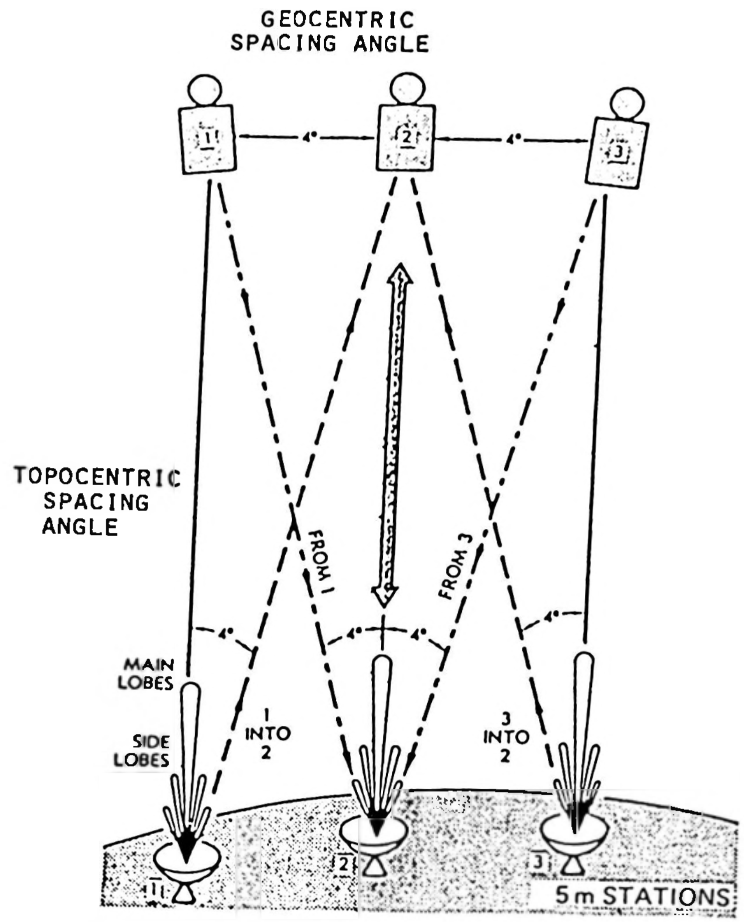
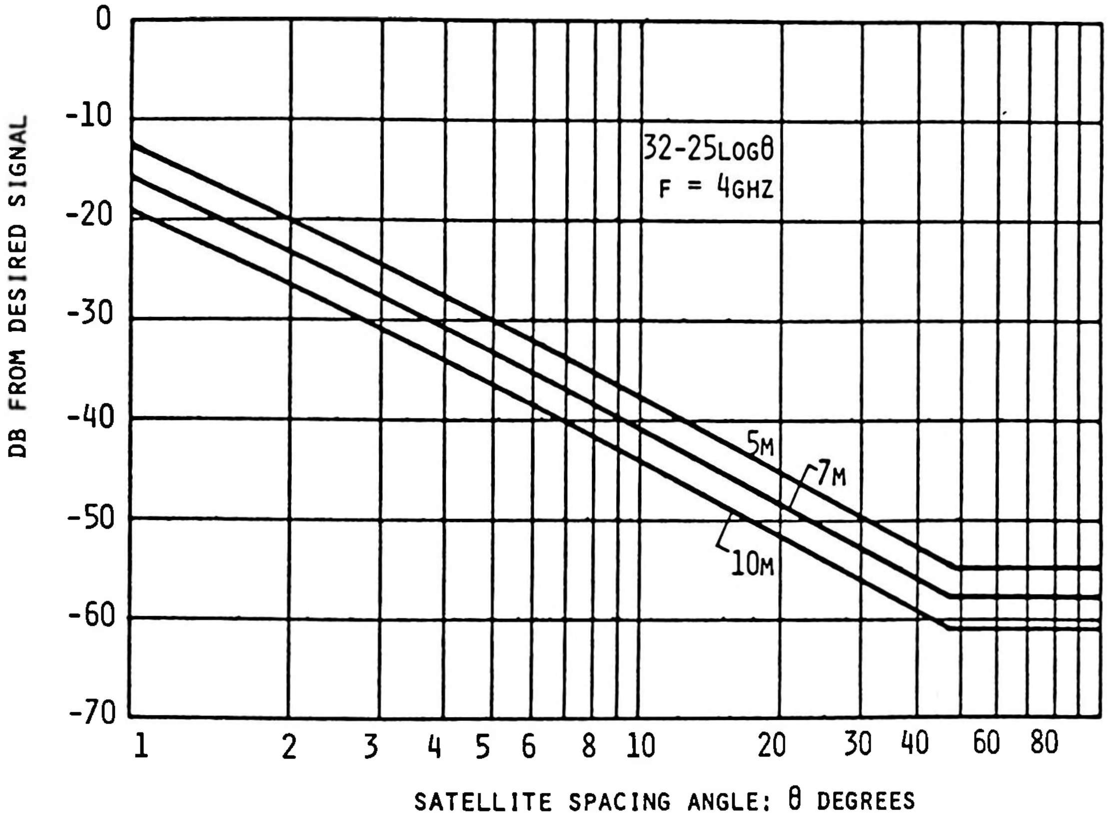


FIGURE 1



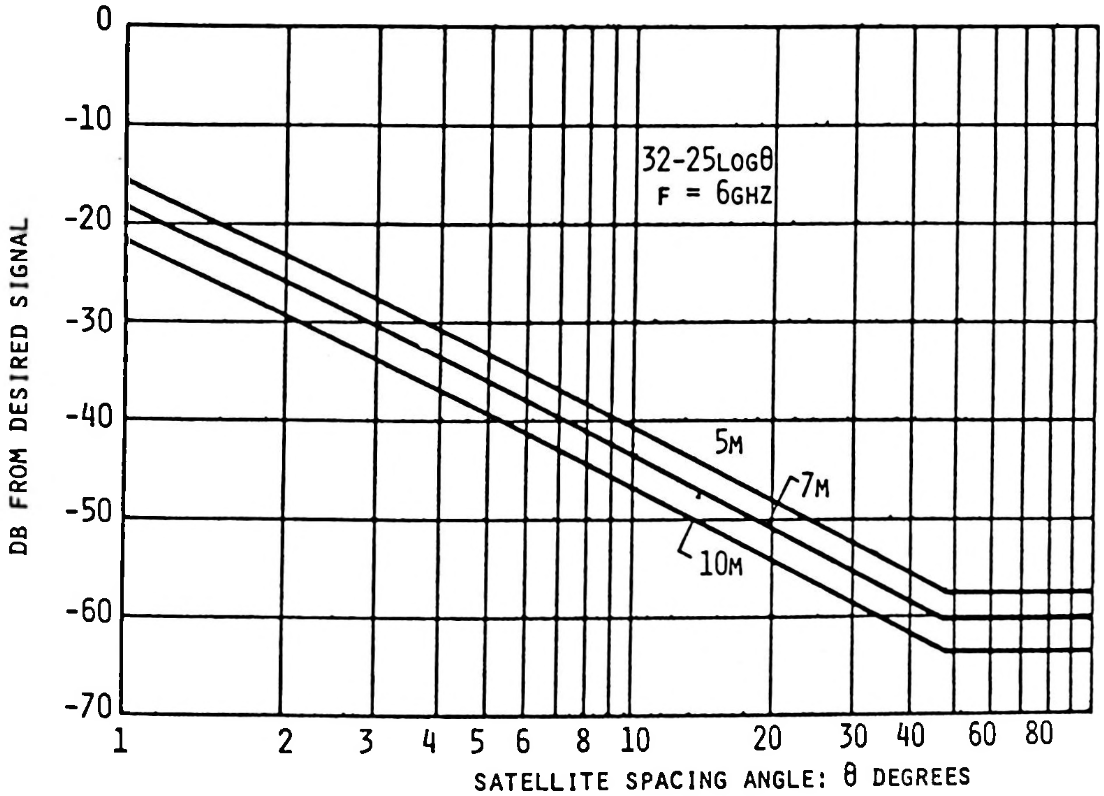
SIGNALS FROM EARTH STATION #2 ARE CONTAMINATED BY BOTH OTHER EARTH STATIONS, AND ARE AGAIN INTERFERED WITH BY SATELLITES #1 AND #3 UPON RETRANSMISSION FROM THE SATELLITE.

FIGURE 3



SIDELOBE INTERFERENCE BECOMES MORE SEVERE AS SPACECRAFT ARE CROWDED CLOSER TOGETHER AND SMALL ANTENNAS ARE USED THAT HAVE LARGER BEAM WIDTHS.

FIGURE 2



SIDELOBE INTERFERENCE BECOMES MORE SEVERE AS SPACECRAFTS ARE CROWDED CLOSER TOGETHER AND SMALL EARTH STATION ANTENNAS ARE USED THAT HAVE LARGER BEAM WIDTHS.

FIGURE 4

FCC NOI DOCKET #81-704
PROPOSES A 3 DB IMPROVEMENT AT SMALL OFF-AXIS ANGLES $1^{0} < \theta \leq 7^{0}$, A PLATEAU BETWEEN THE OLD AND IMPROVED PATTERN AT THE 8 DBI GAIN LEVEL AT ANGLES BETWEEN 7^{0} AND 9.2^{0} OFF-AXIS AND RETAIN THE OLD STANDARD BETWEEN 9.2^{0} AND 48^{0}.

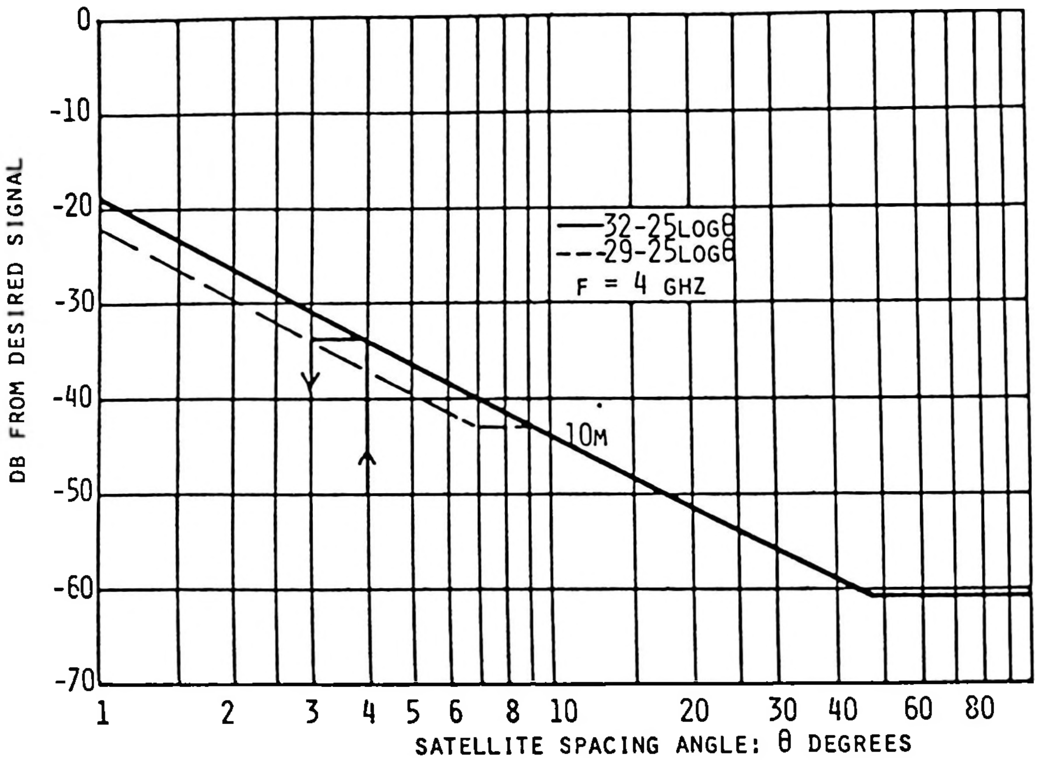
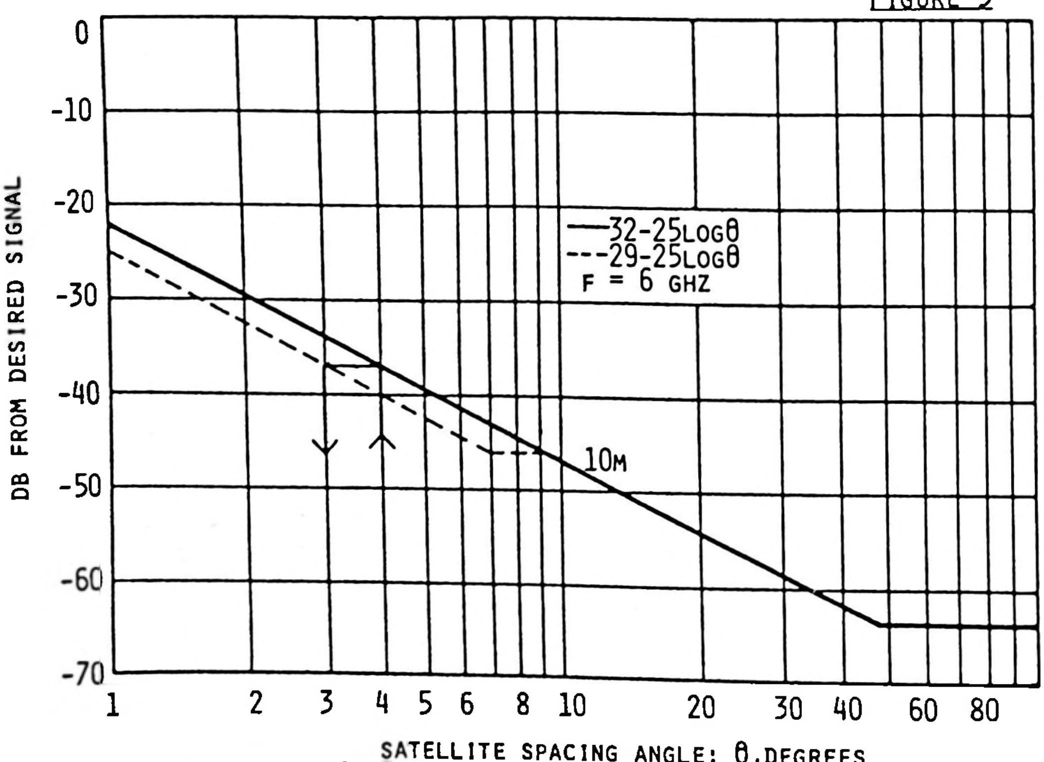


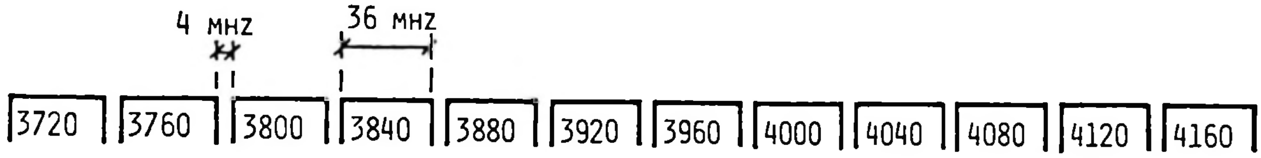
FIGURE 5



FCC NOI DOCKET #81-704
PROPOSES A 3DB IMPROVEMENT AT SMALL OFF-AXIS ANGLES $1^{0} < \theta < 7^{0}$, A PLATEAU BETWEEN THE OLD AND IMPROVED PATTERNS AT 8 DBI GAIN LEVEL AT ANGLES BETWEEN 7^{0} AND 9.2^{0} OFF-AXIS AND RETAIN THE OLD STANDARD BETWEEN 9.2^{0} AND 48^{0}.

FIGURE 6

DOWNLINK: HORIZONTAL POLARIZATION CENTER FREQUENCIES IN MHZ

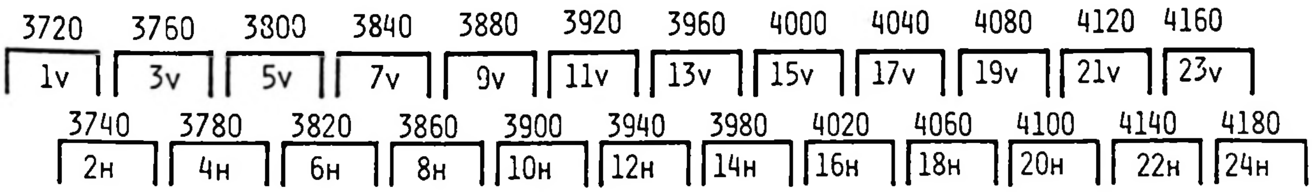


UPLINK: VERTICAL POLARIZATION CENTER FREQUENCIES IN MHZ

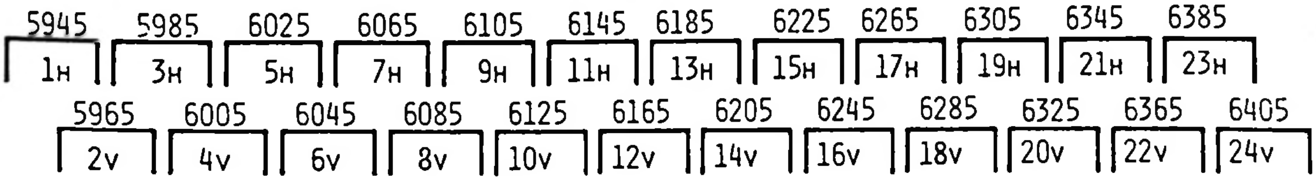
FREQUENCY PLAN EARLY WESTAR SATELLITES

FIGURE 7

DOWNLINK CENTER FREQUENCIES, MHZ

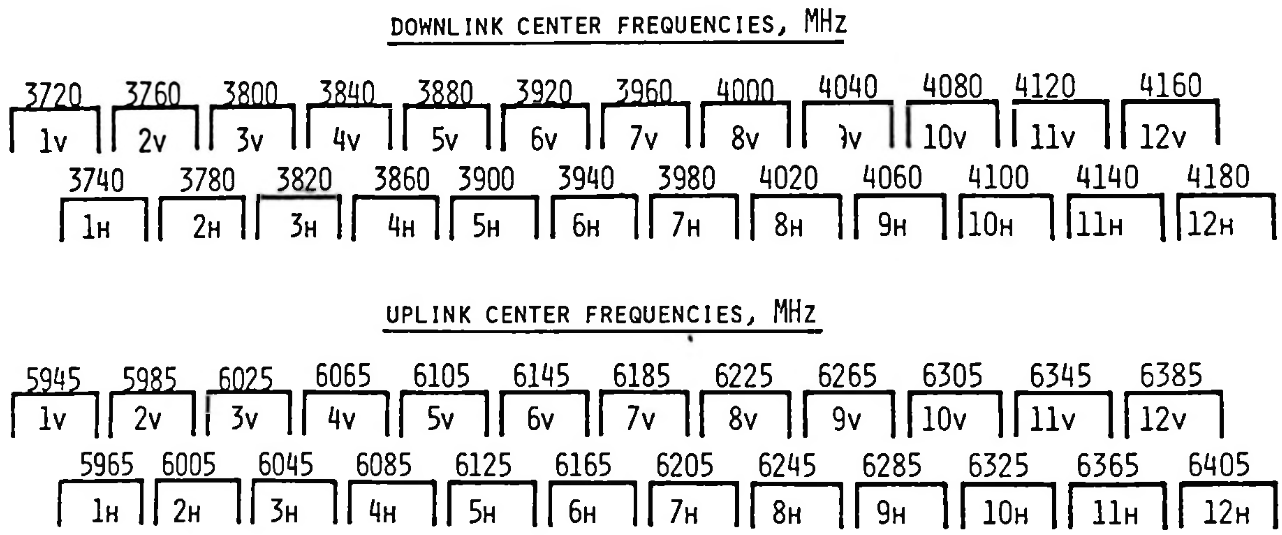


UPLINK CENTER FREQUENCIES, MHZ



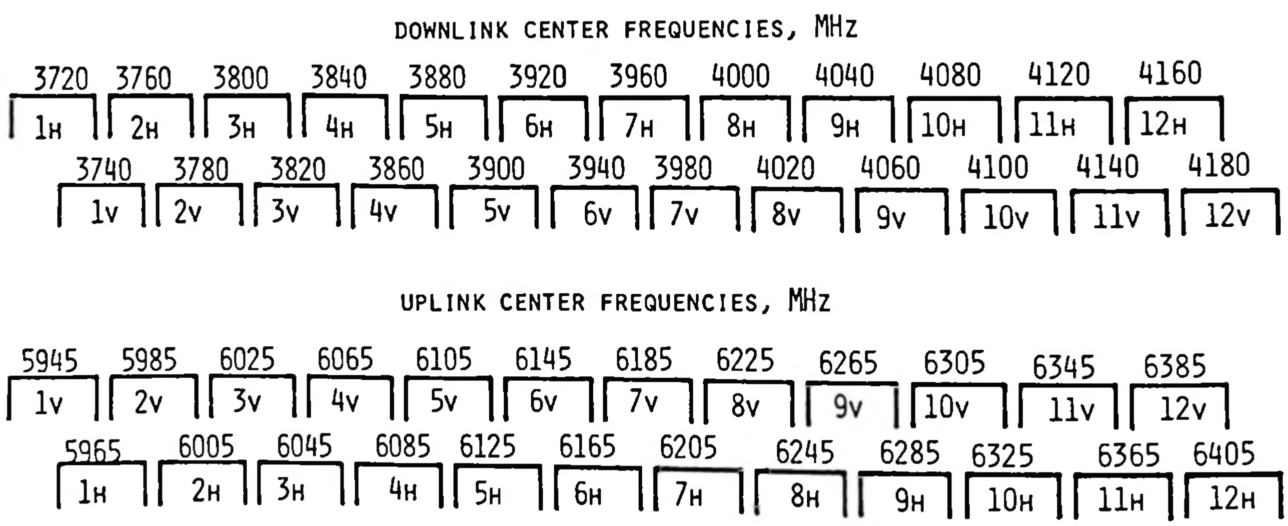
FREQUENCY & POLARIZATION PLAN SATCOM SATELLITES

FIGURE 8



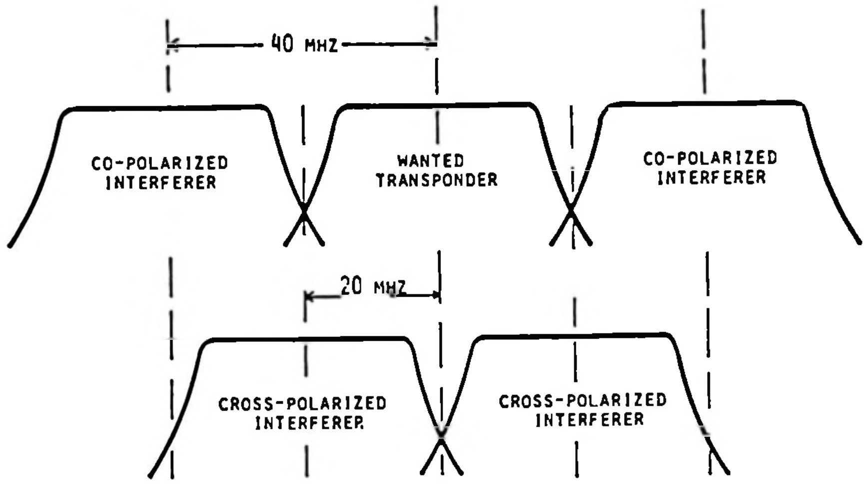
FREQUENCY & POLARIZATION PLAN COMSTAR SATELLITES

FIGURE 9



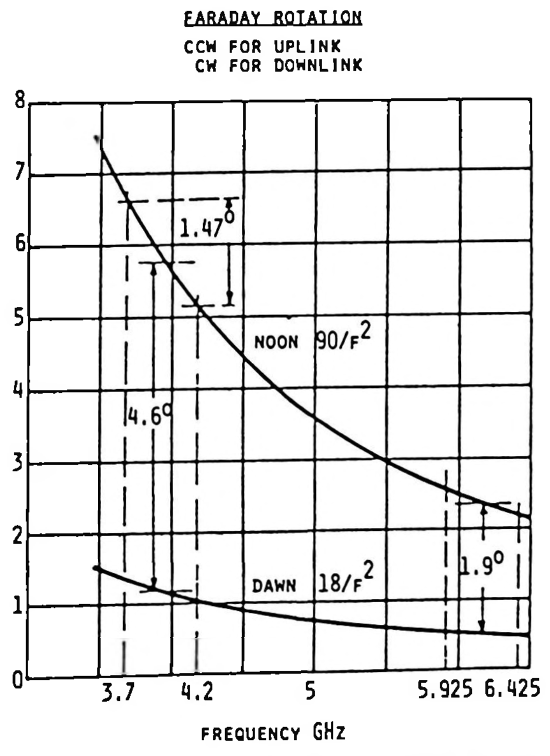
FREQUENCY & POLARIZATION PLAN WESTAR IV & V

FIGURE 10



TYPICAL FREQUENCY AND POLARIZATION PLAN OF FREQUENCY REUSE SATELLITE.

FIGURE 11



FARADAY ROTATION ALTERS THE ORIENTATION OF LINEARLY POLARIZED SIGNALS AS THEY PASS THROUGH THE IONOSPHERE. ROTATION IS MAXIMUM DURING THE DAY. IT SUBSIDES AT COOLER TEMPERATURES JUST BEFORE DAWN.

FIGURE 12

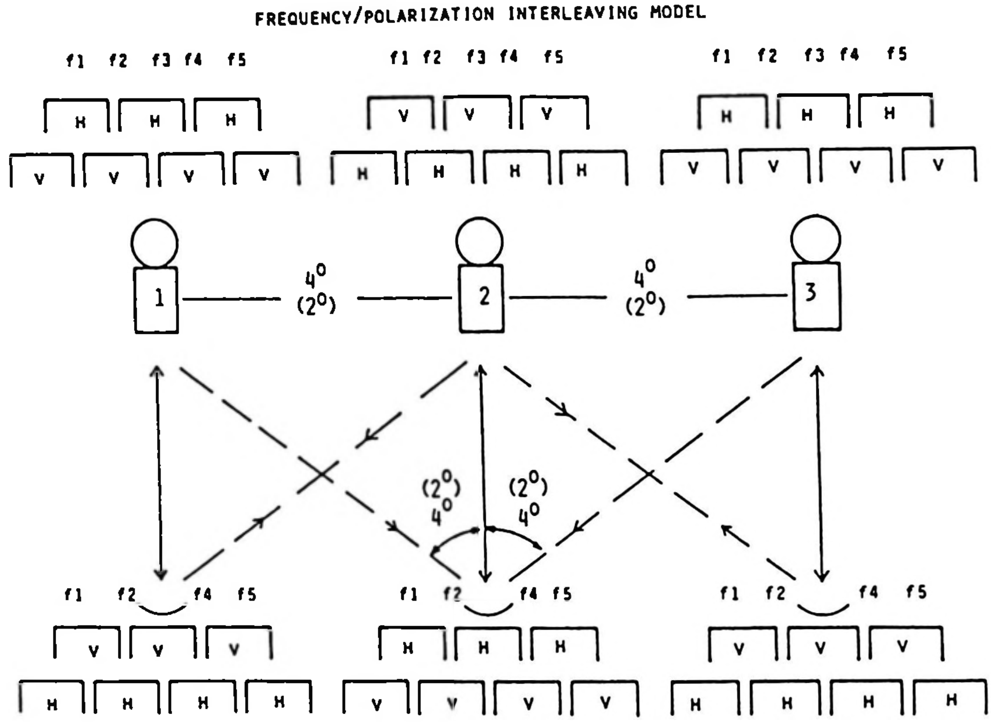
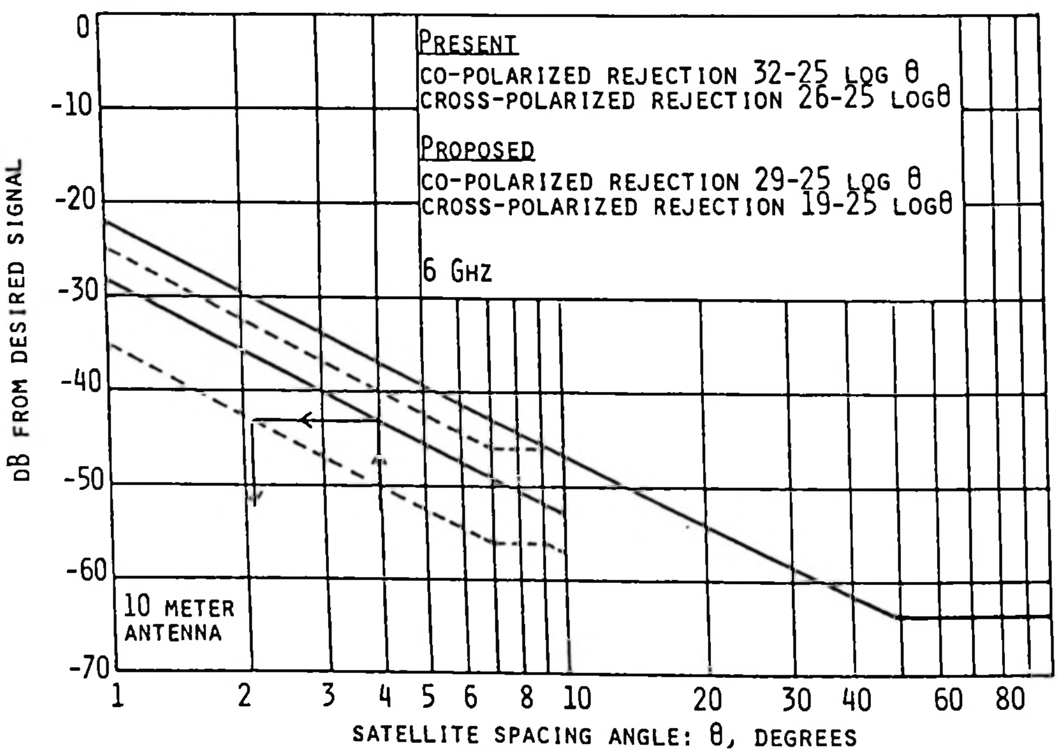
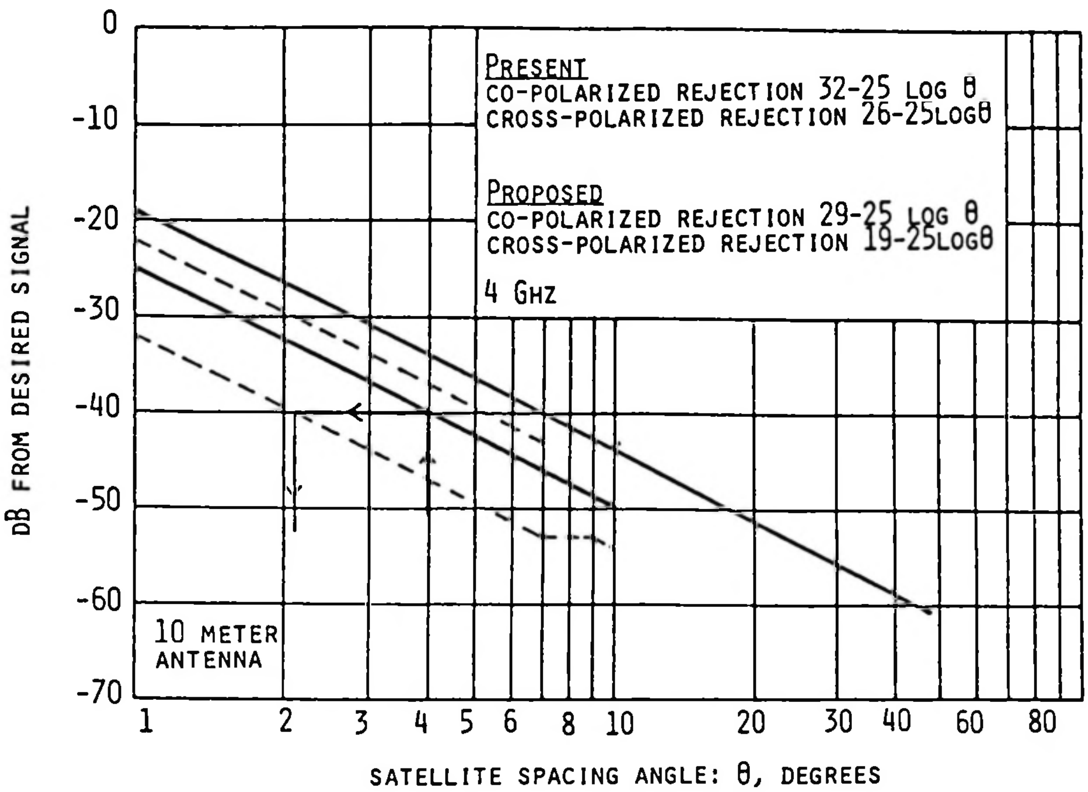


FIGURE 13



PRESENT AND PROPOSED ANTENNA PATTERN ENVELOPES AND EFFECT OF POLARIZATION ISOLATION.

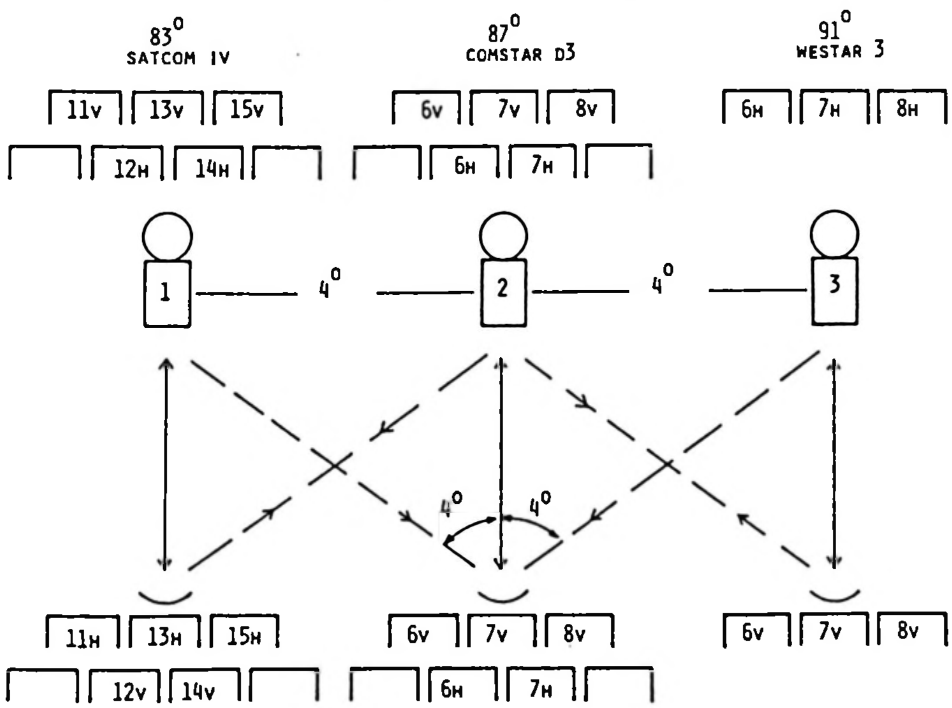
FIGURE 14



PRESENT AND PROPOSED ANTENNA PATTERN ENVELOPES AND EFFECT OF POLARIZATION ISOLATION.

FIGURE 15

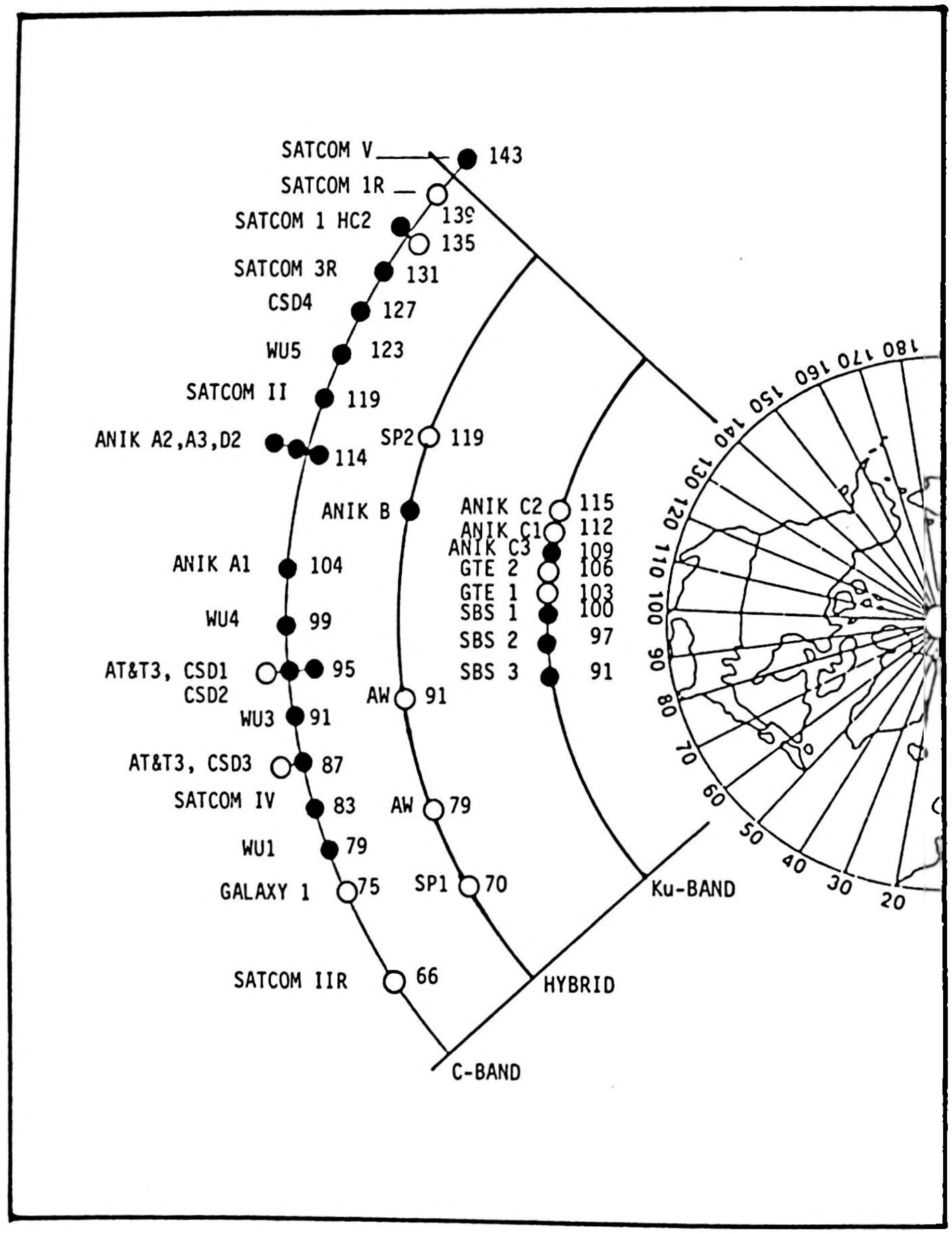
DOWNLINK 4 GHZ FREQUENCY/POLARIZATION PLAN



UPLINK 6 GHZ FREQUENCY/POLARIZATION PLAN

FIGURE 17

FIGURE 16 DOMESTIC COMMUNICATION SATELLITES VIEWING THE U.S.



ANTENNA NOISE TEMPERATURE

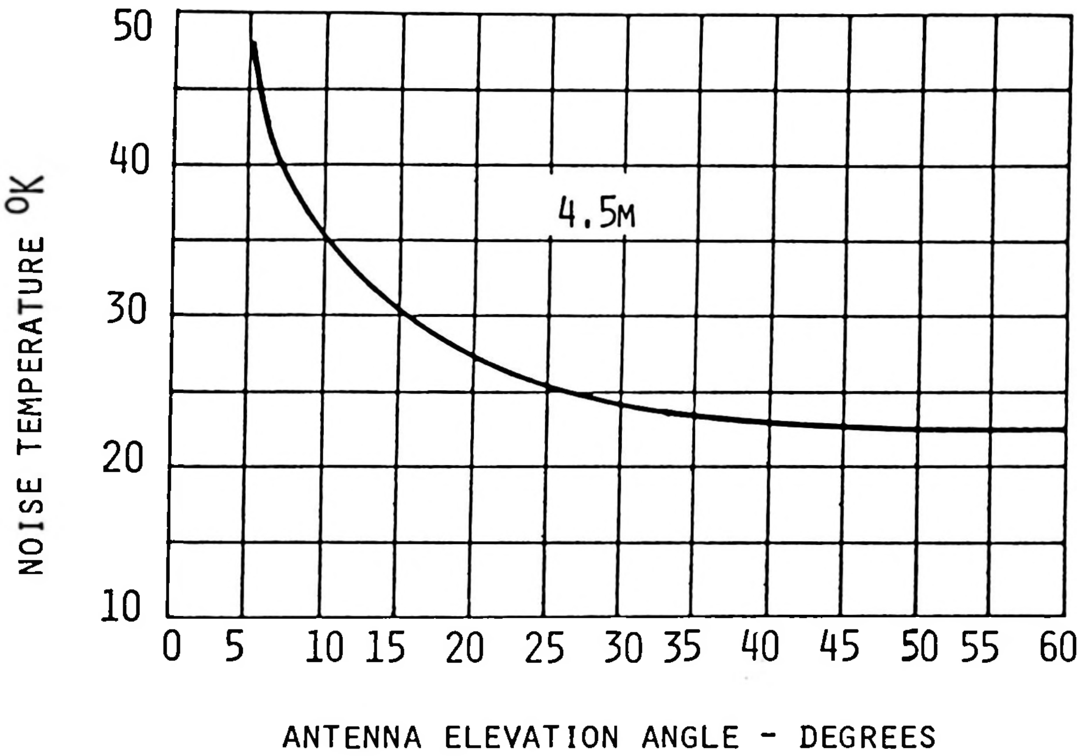


FIGURE 18

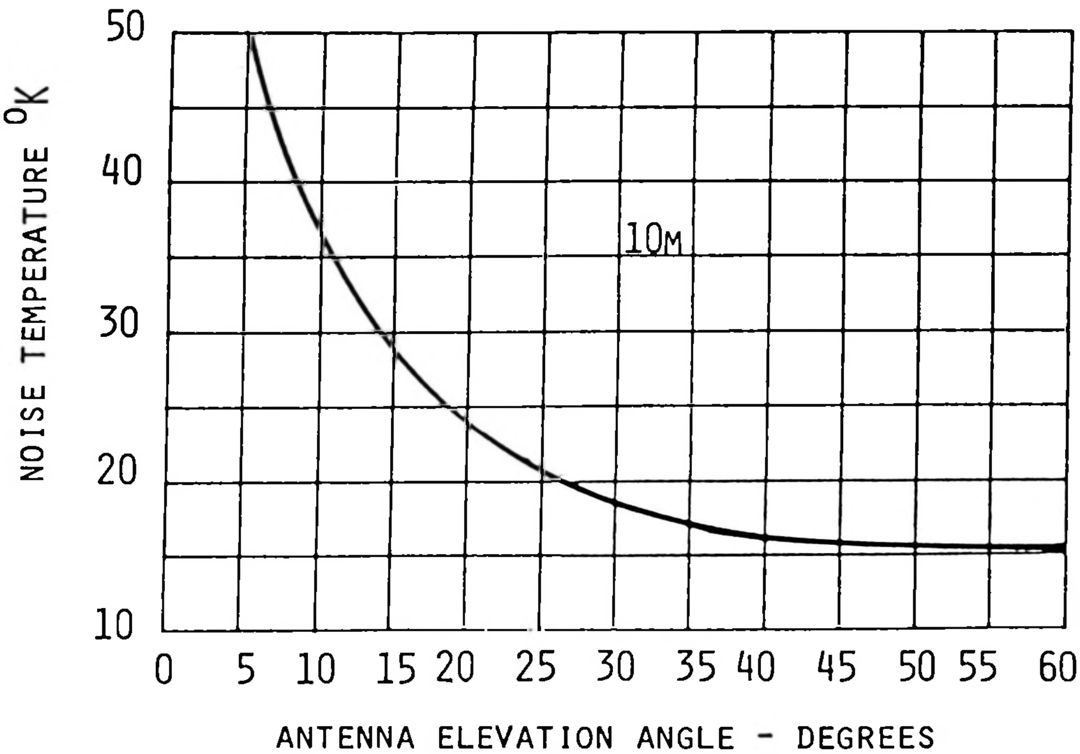


FIGURE 19

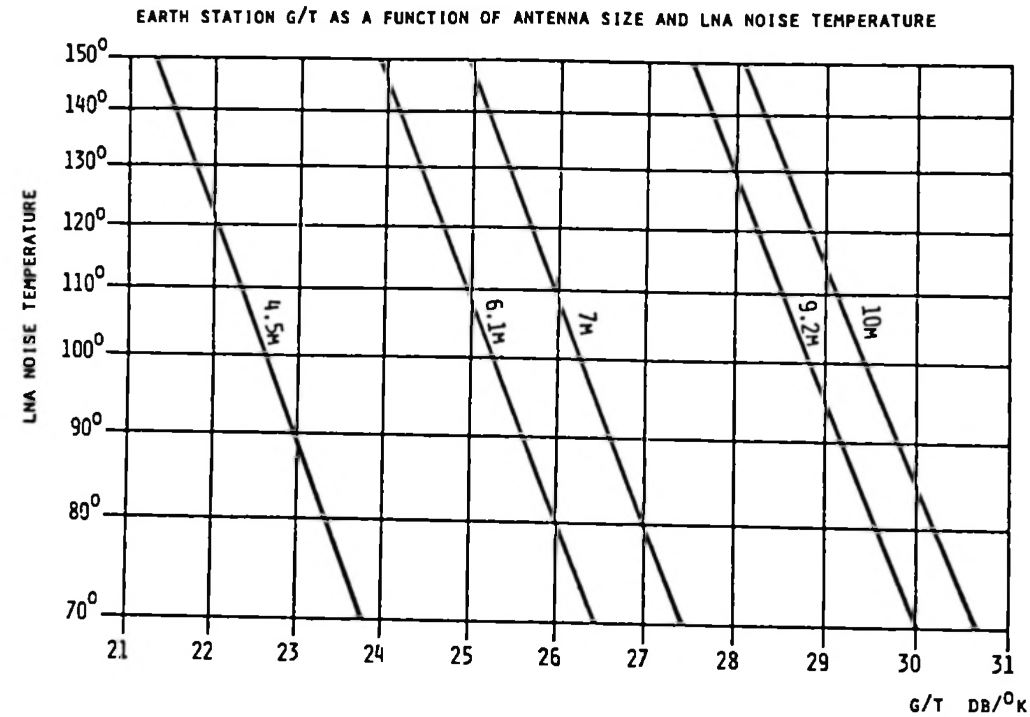


FIGURE 20

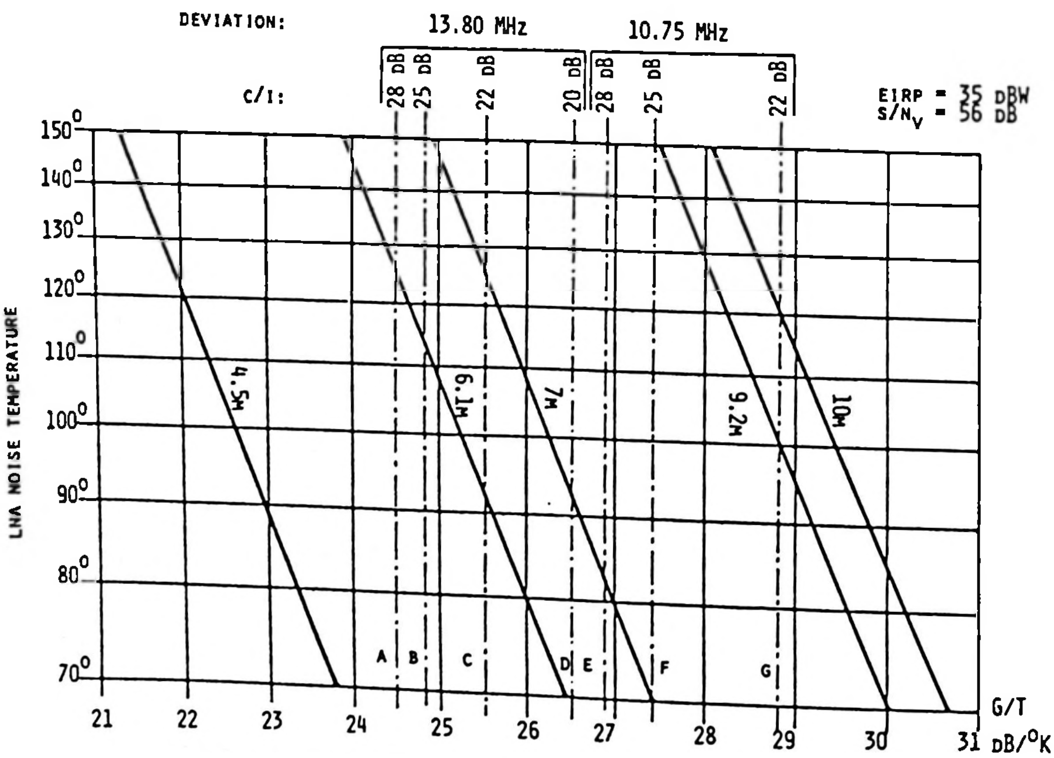


FIGURE 21

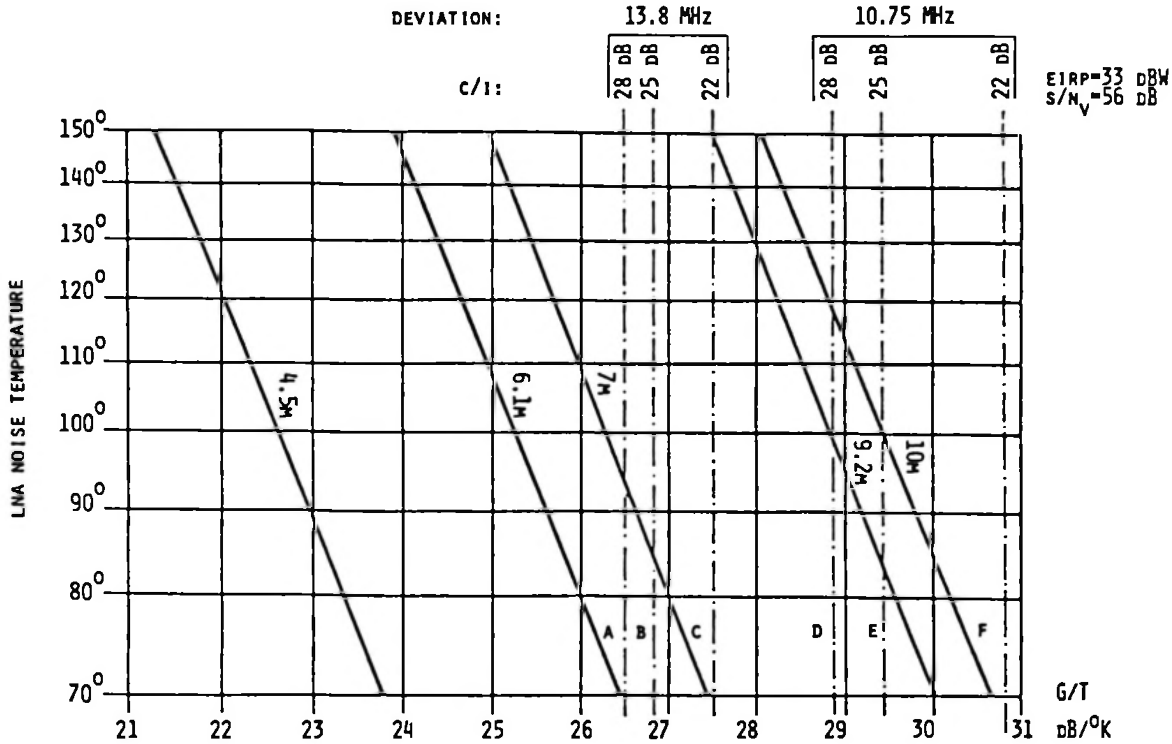


FIGURE 22

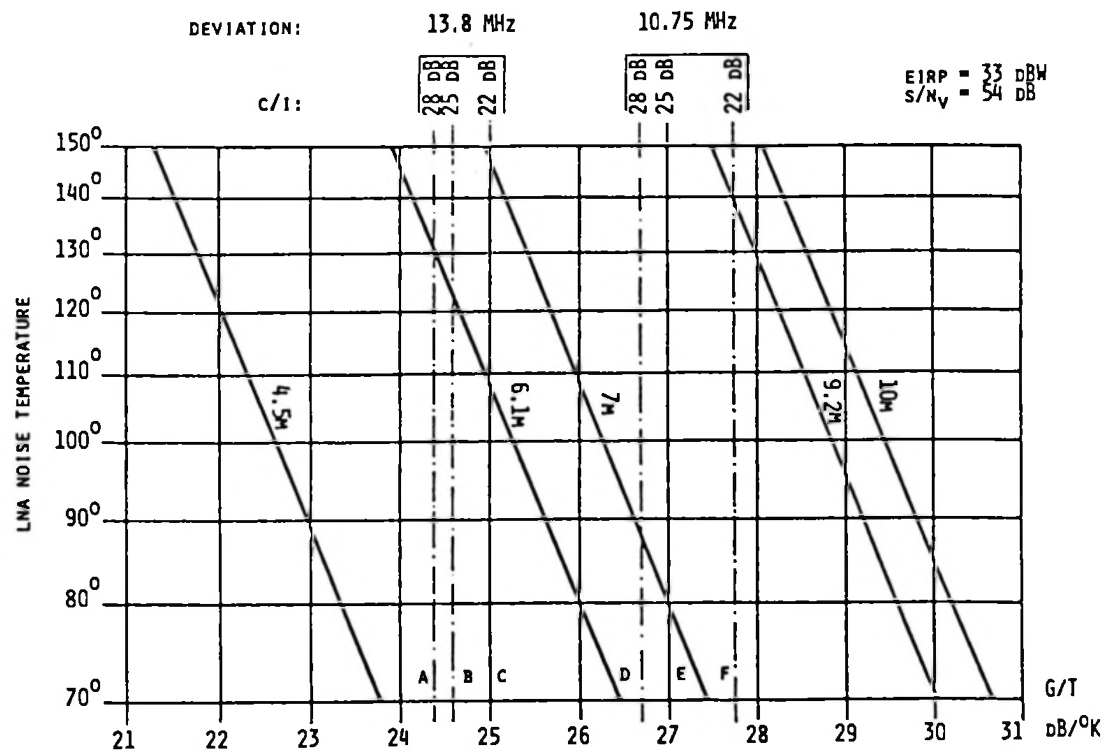


FIGURE 23

DESIGN CRITERIA FOR MULTI-STATION COMBINING SYSTEMS

Spencer J. Smith	Robert R. Weirather
Vice Pres. Engineering	Consultant, Adv. Dev.
Dielectric Communications	Harris Broadcast Division

INTRODUCTION

Section I

The requirements of multi-station FM systems has in the last few years reached the point where it is convenient or cost effective to combine one or two stations into a single antenna and technical expertise has reached the level where dozens or more stations can be combined for efficient performance and be very cost effective.

This paper will go through the standard combining techniques used for multi-station systems as well as the cost impact of such multi-station installations. Data will be presented from the nine station multiplexer built for the Senior Road Tower Group in Houston. Performance in both the R.F. and the audio response of such a system will be analyzed in view of the more strenuous requirements that the newer modulation schemes require. Techniques are examined which allow the handling of these newer modulation schemes within constant resistance networks of both the band reject and band pass variety. In addition, correlation will be given by the measured R.F. response, i.e. insertion loss and group delay vs. the audio response for these various modulation sequences.

TYPES OF MULTIPLEXERS

There are two general categories which are standard today in the U.S. for combining R.F. signals at different frequencies. These are: branch type, (star connected), and balanced type, (constant resistance type). In general, multi-station multiplexing of more than two or three stations dictate the balance type module construction from a performance standpoint. These will be the only type of module examined in this paper.

There are two general styles of balanced multiplexing: one uses a module centered around band pass filters and the other uses modules centered around band reject filters. Each type will be examined and the pros and cons discussed.

In a multiplexer, the common practice is to combine either of these modules in a series of parallel combinations with each module tuned to one transmitter frequency and the requirement being that an individual module be supplied for each transmitter frequency required. In the pursuit of consistency and common parts, a multiplexer is normally constructed of all one type of module, either band pass or band reject. However, this is not necessarily a requirement in most multiplexers.

THE MODULE

The balanced module consists of two hybrids, usually $\pi/2$, a reject load, and two sets of cavities mounted between the hybrids as shown in Figure 1.

In the band reject case, the signal entering Port 1 is divided equally in the first hybrid and travels to the reject filters which are tuned to that frequency. In this case, the cavities present cross to a short circuit for this signal and the reflections return to the first hybrid and recombine at Port 2. This leaves the second hybrid isolated completely by the notch depth of the cavities. A signal of a different frequency can enter the second hybrid through Port 3, is divided and passes to the first hybrid since the notch cavities are transparent at other frequencies. Thus a group of these modules connected in series will allow the culmination of a multiple of R.F. signals in different frequencies.

In the band pass module, the signal that enters Port 3 of the second hybrid sees the two filters in the pass band and are thus transparent to the signal, recombines in the first hybrid and exits Port 2 of that hybrid. Frequencies other than that of the band pass filters can enter Port 1 of the first hybrid, see the band pass cavities as short circuits and are reflected to Port 2 of the first hybrid. A group of these modules connected in parallel will operate in a very similar manner as a group of the band reject modules connected in series.

REJECT MODULE MULTIPLEXER

A multiplexer composed of reject modules would be considered a series connected unit as shown in Figure 2. Of late, the accepted procedure in this configuration is to leave the broadband port available but terminated in module #1. Each transmitter than has its own module to inject its particular frequency into the system. Therefore, N transmitters require N modules. The rationale behind this is that in the event of any module failure, the failed module could be by-passed and the transmitter of that module could use the broad band port at a slight reduction in specifications. Also the use of the broad band port for a permanent transmitter would not allow similar protection to that port as all others in the system; this is probably debatable.

The design of this type of multiplexer is usually done in either decending or ascending frequency from module #1 through N. This allows some liberties to be taken with the reject filters in each module. The critical part of this type of module for the newer modulation schemes is the use of more than one reject filter cavity in each arm between the hybrids (refer to Figure 3).

The responses of a single reject filter are shown in Figure 4. Also shown is how the shape of the reject curve can be adjusted to favor the high side or the low side of the notch frequency.

BANDPASS MODULE MULTIPLEXER

The use of bandpass modules in a multiplexer would result in a parallel connection configuration as shown in Figure 5. As in the band reject concept, each transmitter would have its own module to inject its particular frequency into the chain. This configuration would also allow for a broad band input port if N modules were used for N transmitters. The final complexity of each module, depending on specifications, will in general balance out between the band reject and band pass configurations. This is to say, for a given rejection, the basic quantity of cavities for each module, either between the hybrids or on the narrow band input, will be approximately the same for either case.

BAND REJECT vs. BAND PASS

In the initial design of a multiplexer a major decision is band reject or band pass. In general, a band pass has a greater preponderance of positive features, save one. All else equal, band reject can work with closer frequency spacing, i.e. 800kHz. This is particularly true as the new electrical requirements develop for each narrow band input (refer to Table I).

Bandwidth	$f_o \pm$ 200kHz
Insertion Loss	min. at f_o + 1.5dB at $\pm$ 200kHz
Group Delay	$\pm$ 100 nsec $f_o \pm$ 200kHz
VSWR	1.05:1 at f_o 1.15:1 at $\pm$ 200kHz
Isolation N/B to N/B	50dB to 80dB min. at f_o

NARROW BAND SPECIFICATIONS

TABLE 1

These, combined with features such as natural air convection cooling, by-pass capability, and modular construction, lead to a new level of combiner performance and sophistication.

COSTS

The bottom line in most technical ventures is cost. Is it the least expensive alternative and will it pay for itself. The fixed costs in all installations are the transmitters, and for all practical purposes the tower, i.e. some price increase for higher weight and windload. This reduces the problem down to antenna cost, transmitter line cost, and combiner cost. If we let a single station cost represent 100 percent per station, then how does this amount change as the number of stations increase? The main assumption is that you start with a panel antenna of normal specifications. The transmission line cost increases proportional to size, however, due to better insertion loss, larger than required line is used for single stations. In order to examine the problem, let's put out some concrete specs:

Transmitter power	20KW each
Smallest transmission line	3-1/8"
Antenna	8-bay omni circularly polarized
Tower height	1000'

The approximate cost per transmitter in a multi-station installation is shown in Figure 6. Even though these are relative numbers, they show a definite decrease in cost per station as the number of stations increase. The overall magnitude of the curve can move up and down along the cost axis depending on component selection to put the system together. However, I believe the slope of the curve will remain relatively constant as shown in Figure 6.

SENIOR ROAD TOWER GROUP MULTIPLEXER

The question then comes down to what has been done, or what can be done. The following set of data shown in Figures 7 and 8 and in Table 2 represent measured data on the Senior Road multiplexer from a particular narrow band input to the output of the complete multiplexer. This was done at two different narrow band inputs, one representing a unit that is working 800KC away from a sister station, and the other showing a unit that is operating 1.2 megacycles away from its nearest neighbor. All these measurements are made from the narrow band input to the antenna line.

ISOLATION INPUT TO INPUT AT CENTER BAND ONLY

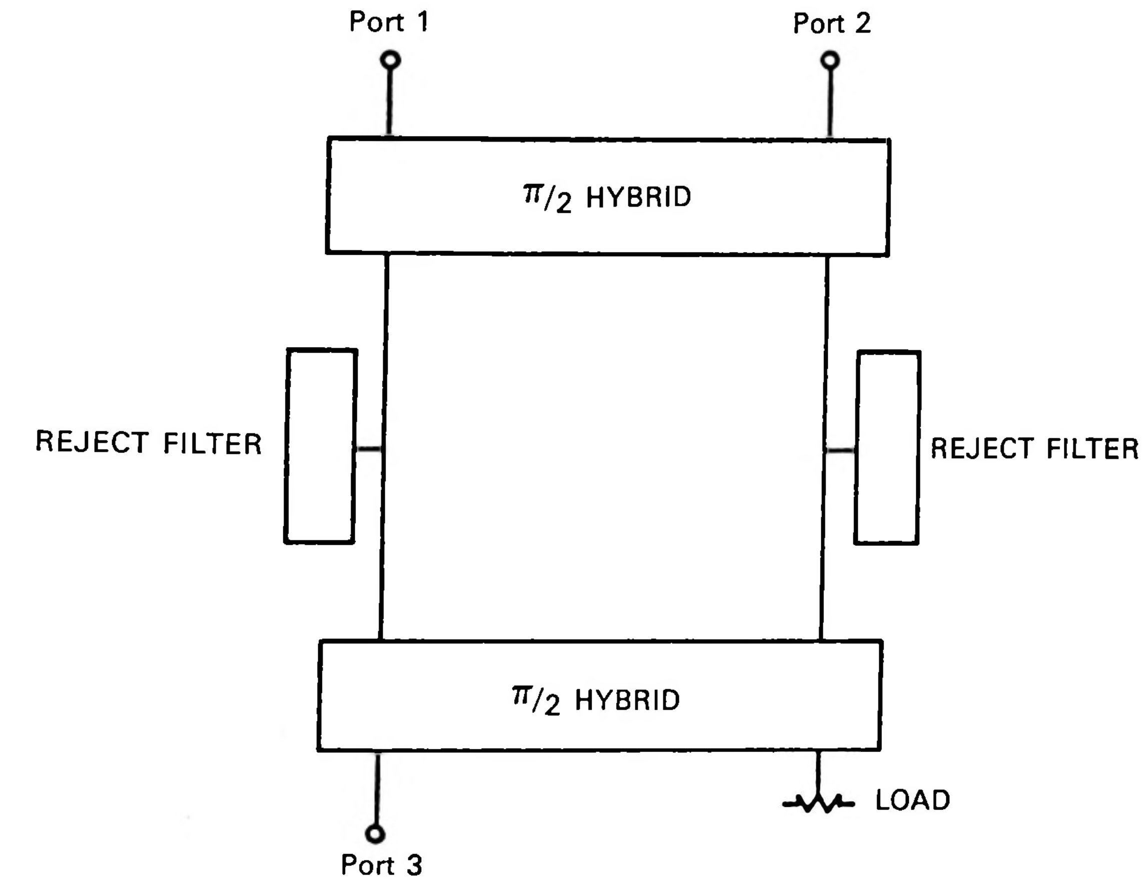
FROM \ TO MODULE	1	2	3	4	5	6	7	8	9
1	x	80	58	66	66	65	82	82	77
2	104	x	77	61	64	61	71	71	85
3	108	102	x	84	60	57	63	64	77
4	104	106	78	x	78	53	58	60	71
5	110	110	100	98	x	72	55	57	70
6	110	110	110	110	108	x	72	51	66
7	110	110	108	108	106	94	x	66	54
8	108	108	106	104	104	104	92	x	51
9	108	108	108	108	108	108	108	108	x

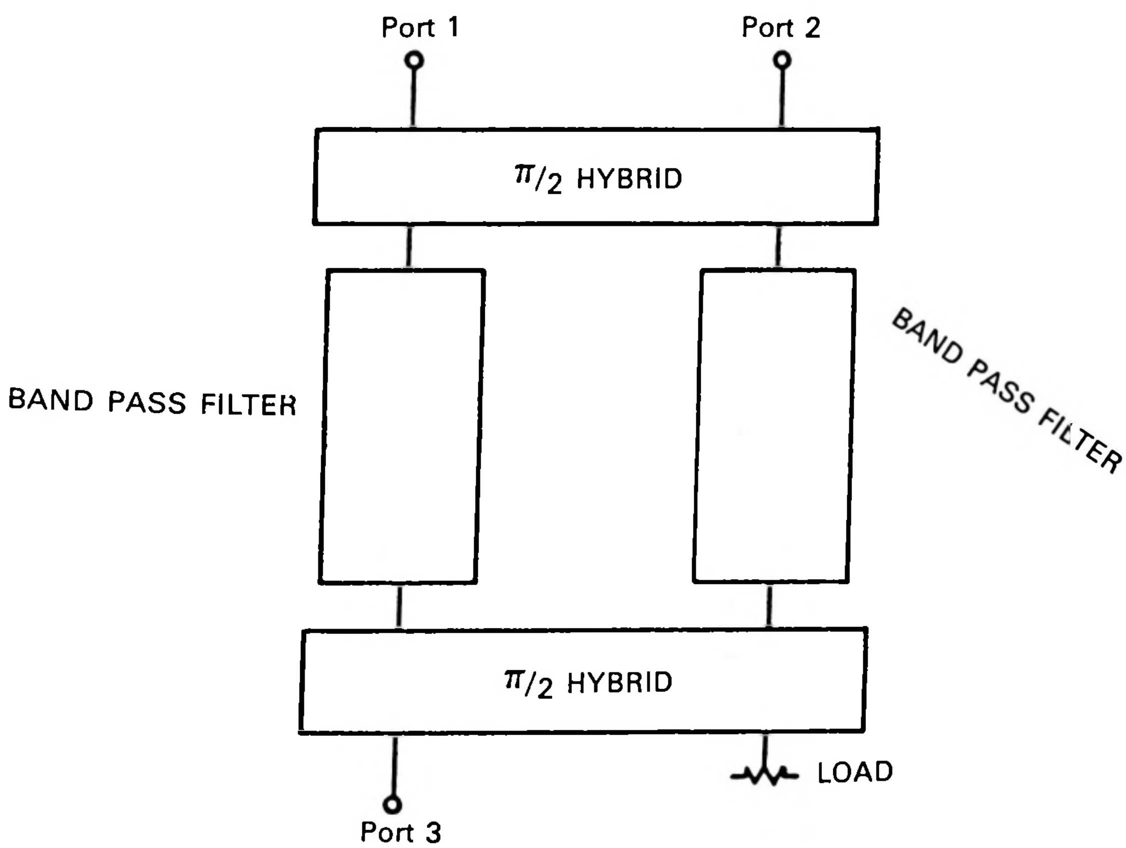
ISOLATION CHARACTERISTICS OF 9-CHANNEL MULTIPLEXER

TABLE 2

The degradation in performance between the two is obvious, however, it should be noted that even the poorer input met all the FM modulation requirements as shown in the second section of this paper. No data is shown for single reject filters in each arm since the performance was obviously so poor that the FM requirements of the unit could not be met. Typically the ±200KC points on insertion loss were down 3dB, and on group delay were down 600 nsec. This would be an unacceptable set of conditions to try to operate. The addition of more cavities in each reject filter assembly could possibly improve the group delay characteristics as well as the insertion loss characteristics on the narrow band inputs. However, no concrete data is yet available on this.

The second part of this paper will discuss how the FM performance is affected by these various RF parameters.

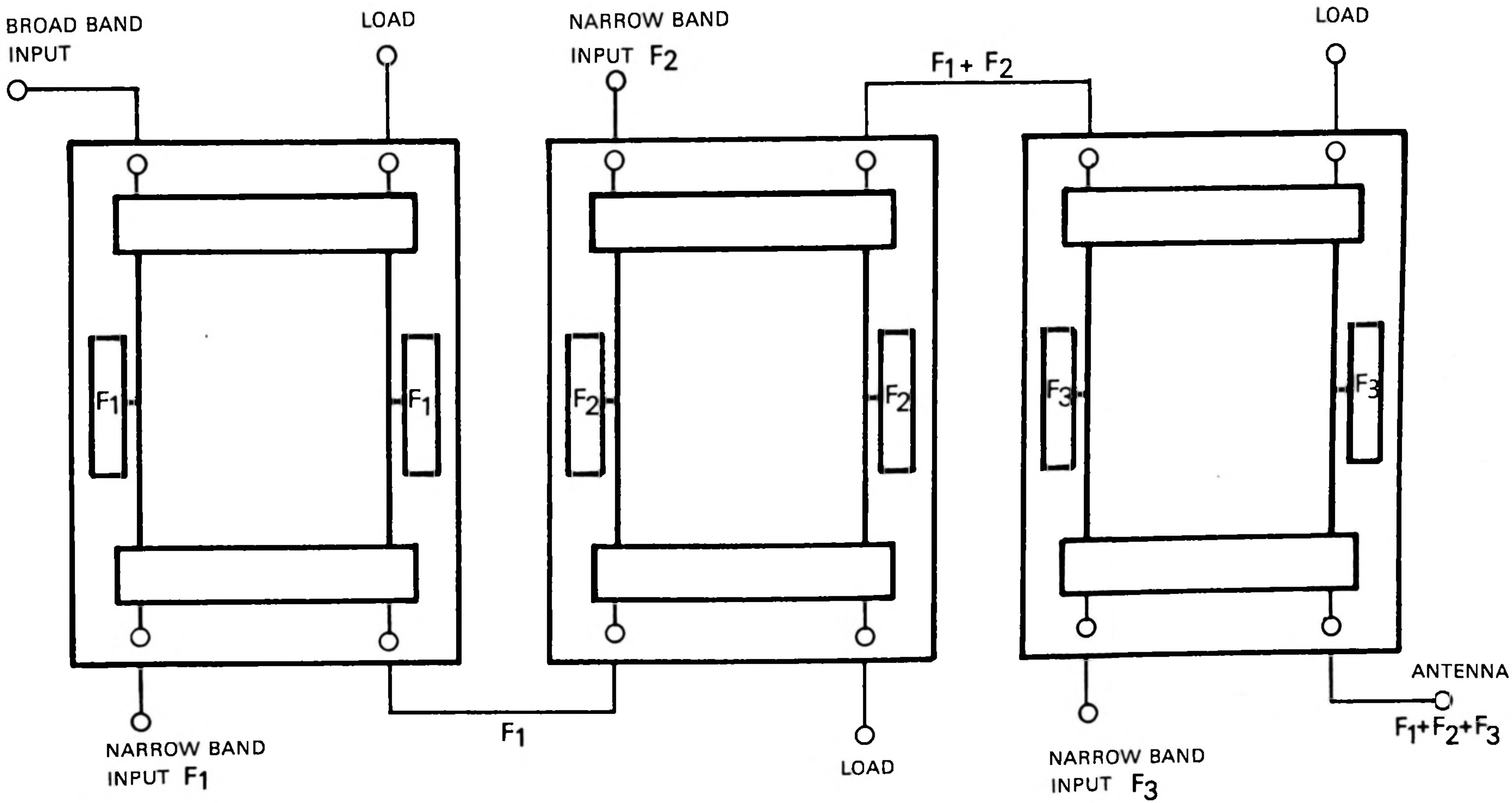




BAND PASS MODULE

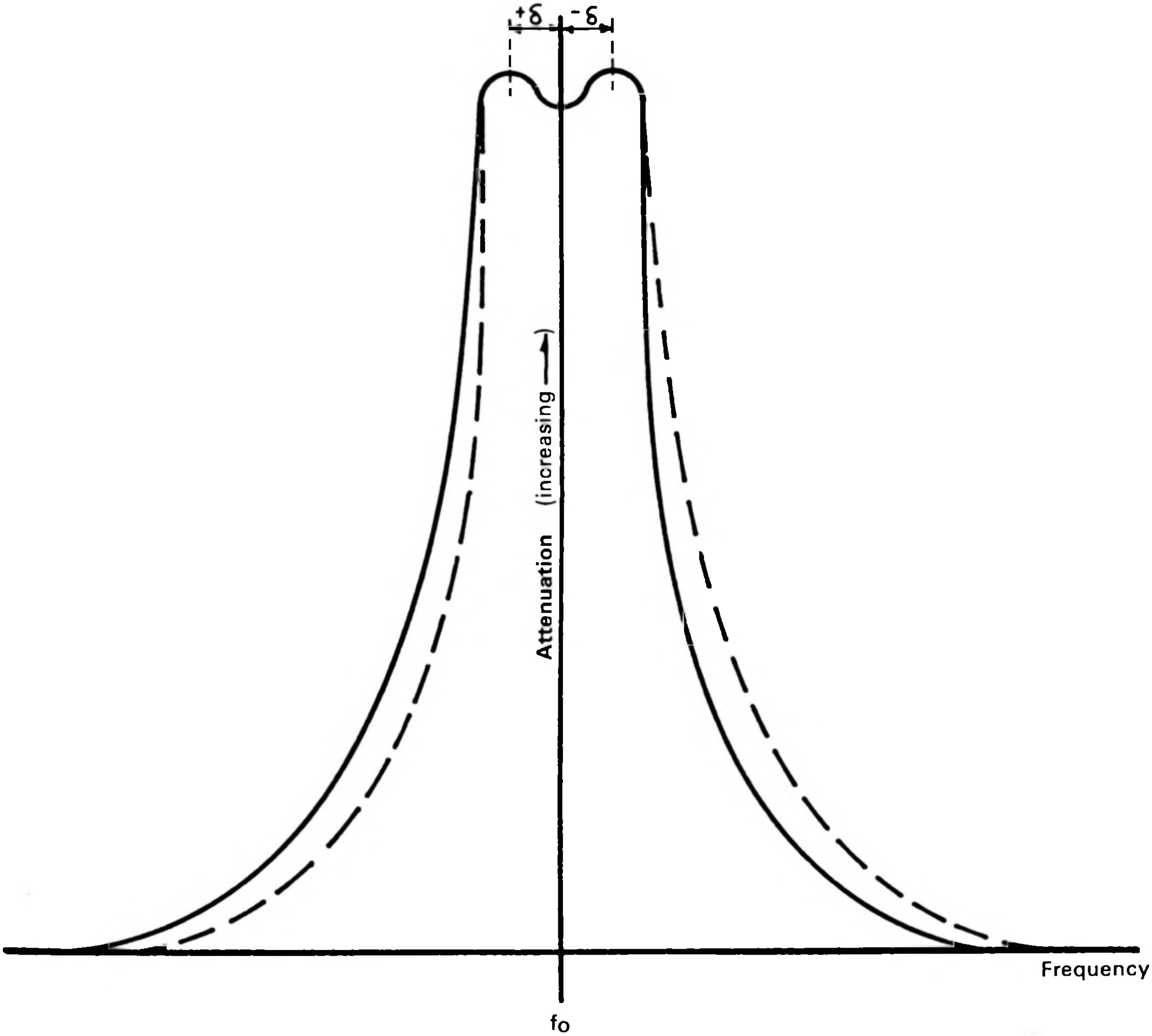
BALANCED MODULES

FIGURE 1



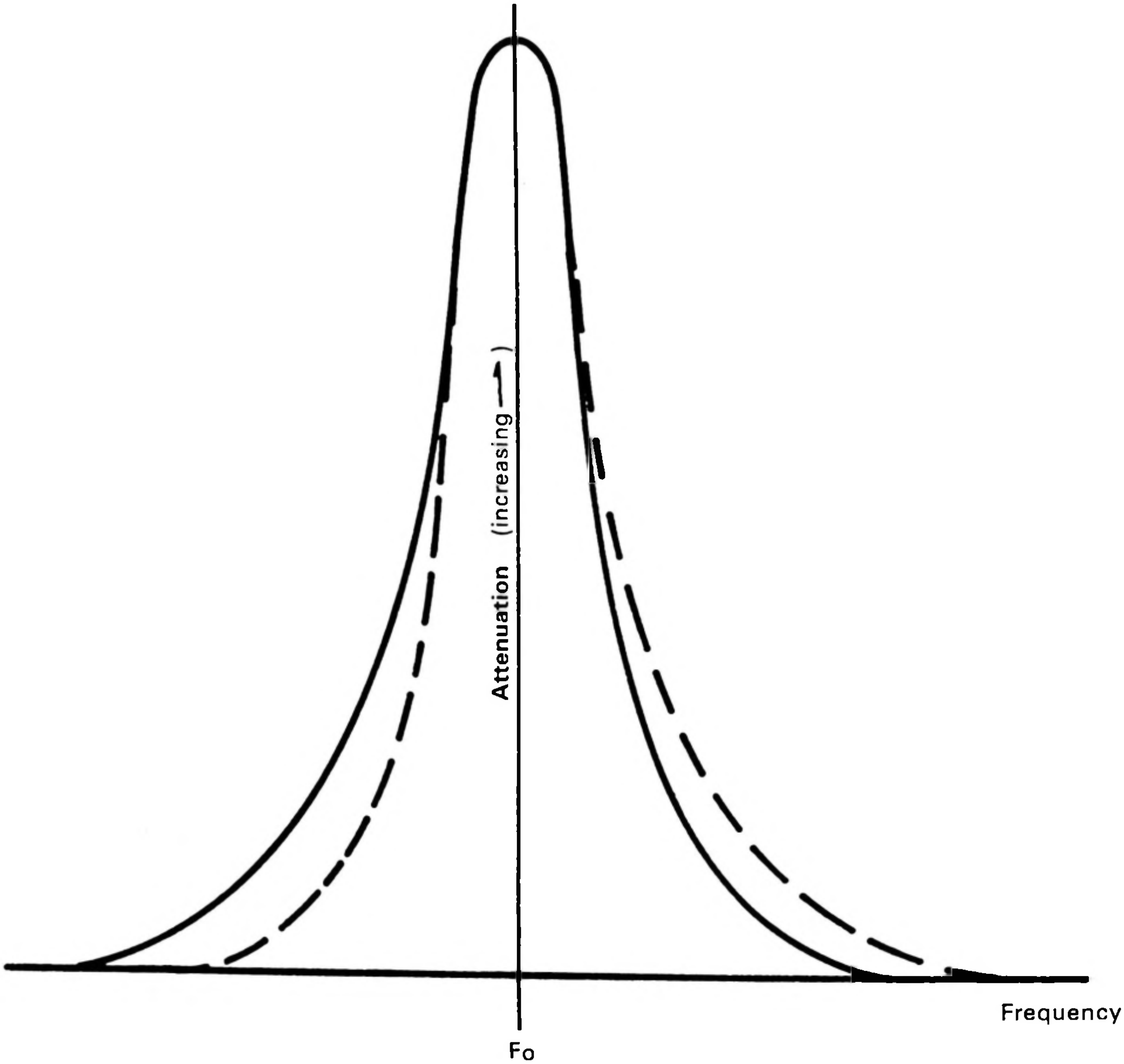
REJECT MODULE MULTIPLEXER

FIGURE 2



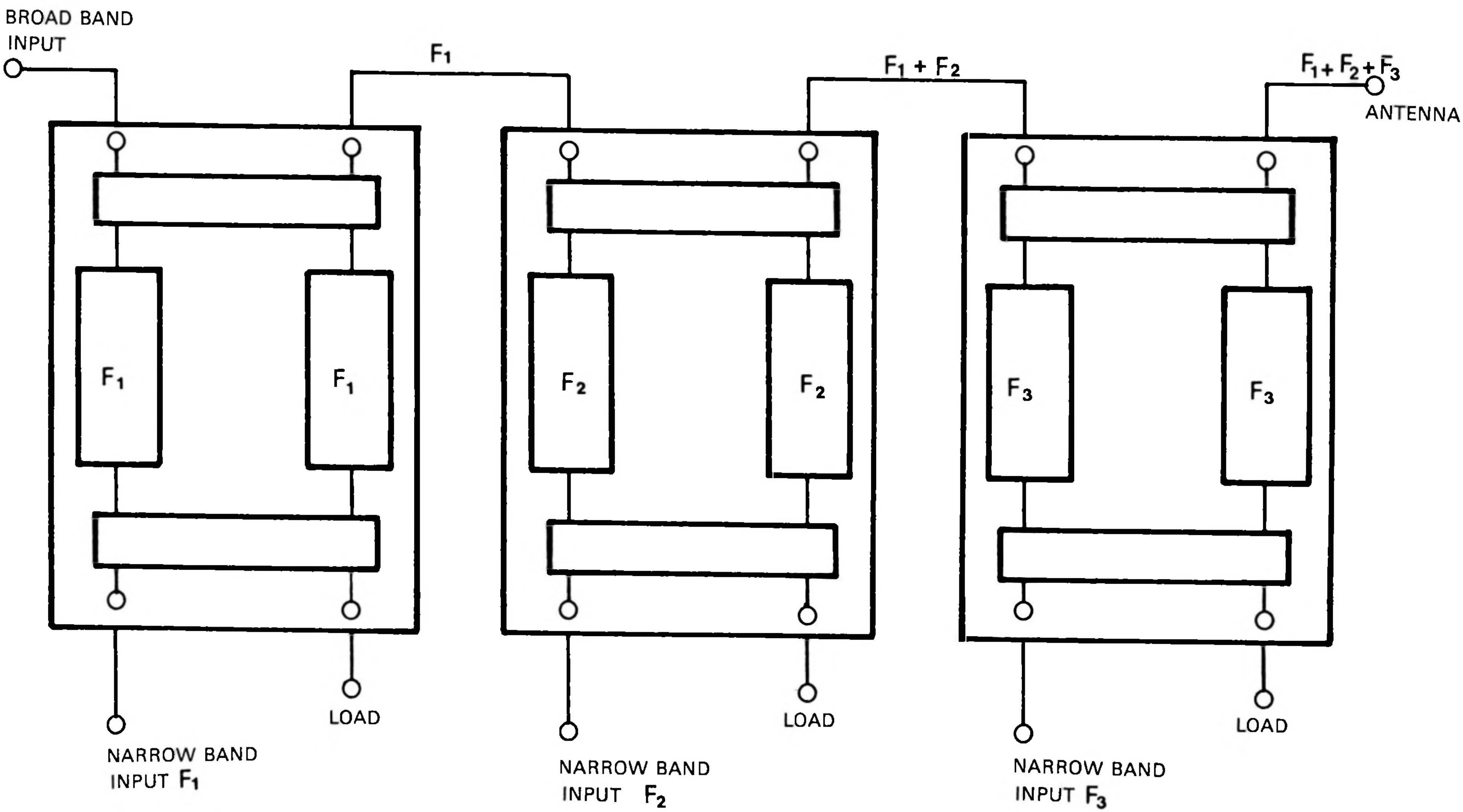
RESPONSE OF TWO CAVITY BAND REJECT

FIGURE 3



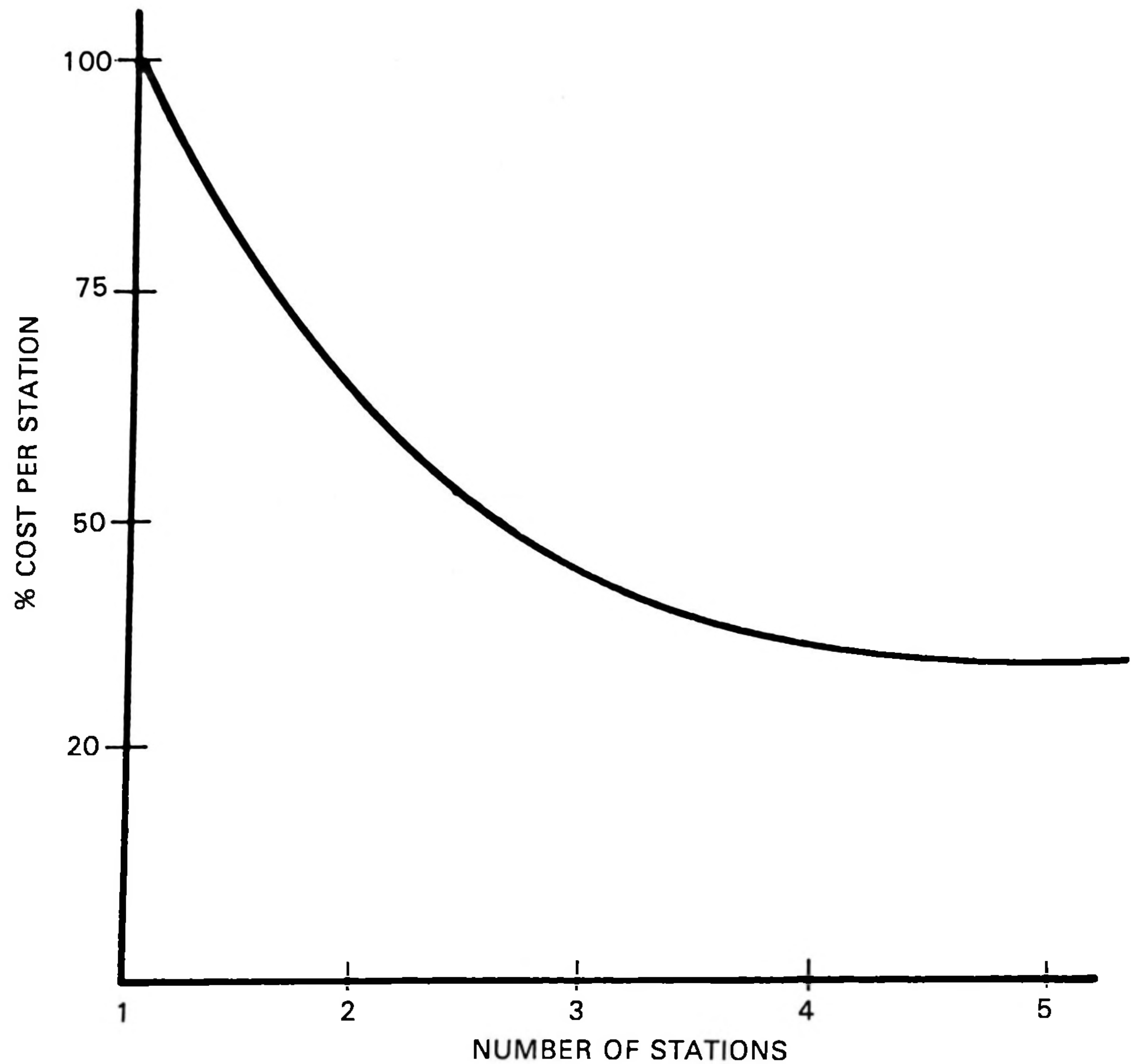
RESPONSE OF SINGLE CAVITY BAND REJECT

FIGURE 4



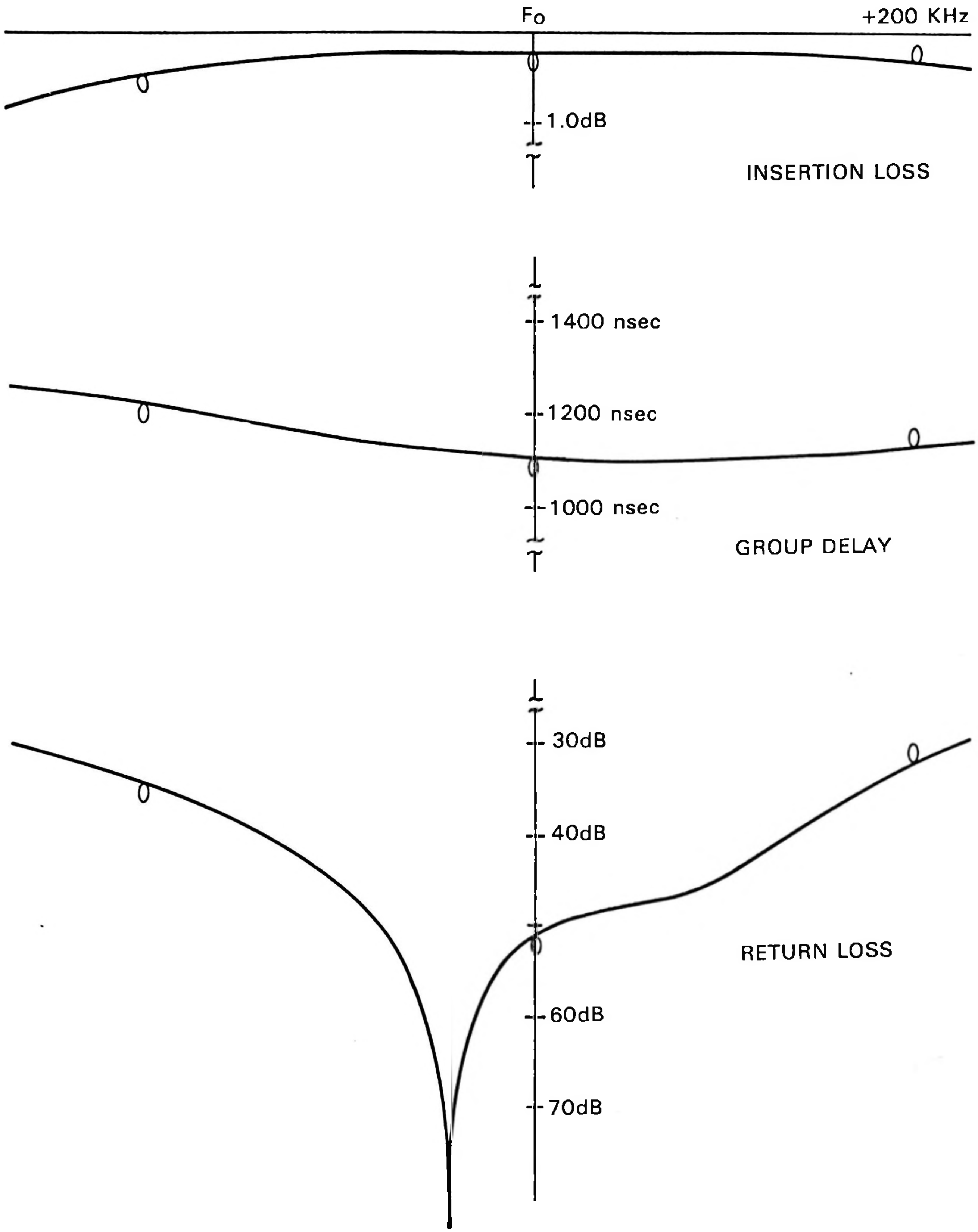
BAND PASS MODULE MULTIPLEXER

FIGURE 5



RELATIVE COST PER STATION

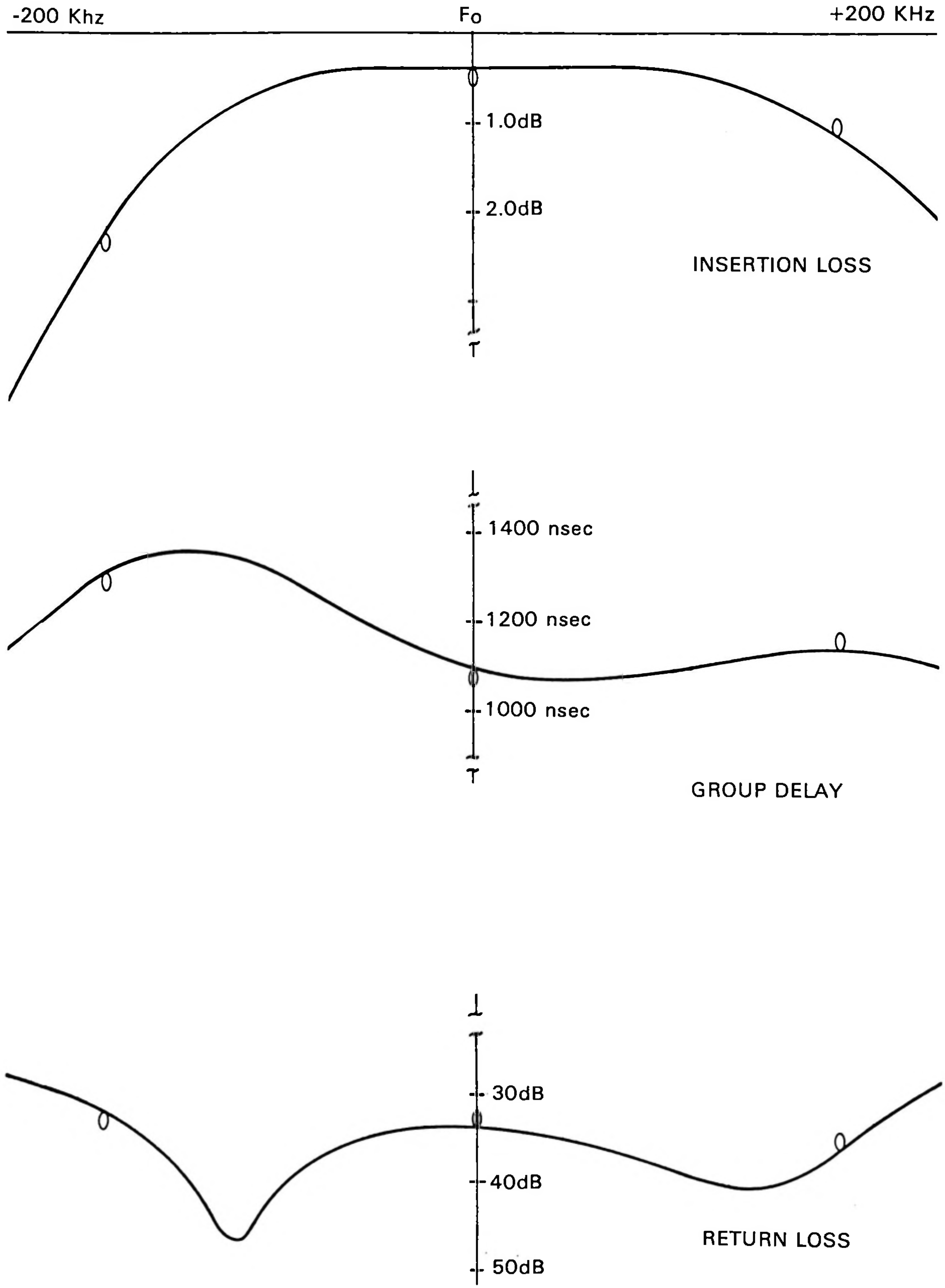
FIGURE 6



NARROW BAND PORT TO ANTENNA PORT CHARACTERISTICS

(1.2 MHz to next frequency)

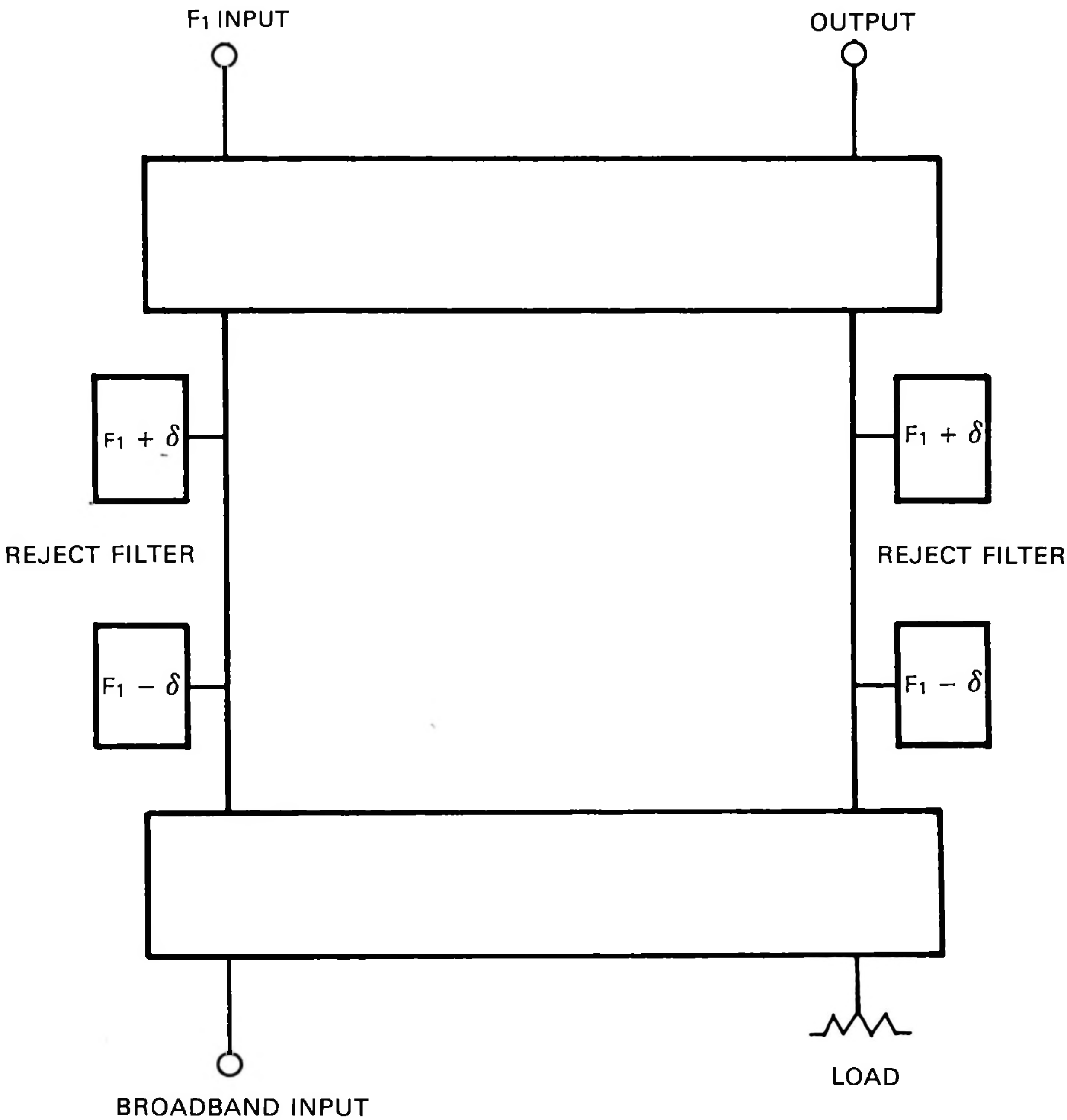
FIGURE 7



NARROW BAND PORT TO ANTENNA PORT CHARACTERISTICS

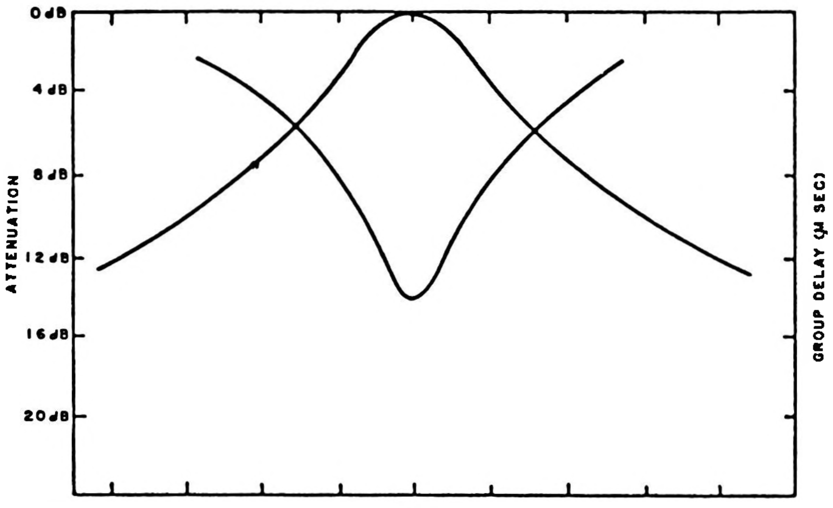
(.80 MHz to next frequency)

FIGURE 8

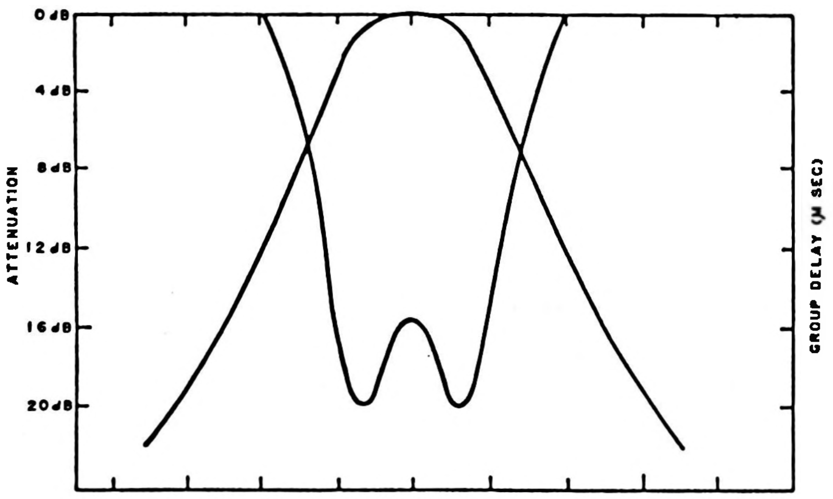


BALANCED DUAL CAVITY REJECT FILTER MODULE

FIGURE 9



RESPONSE OF SINGLE POLE FILTER



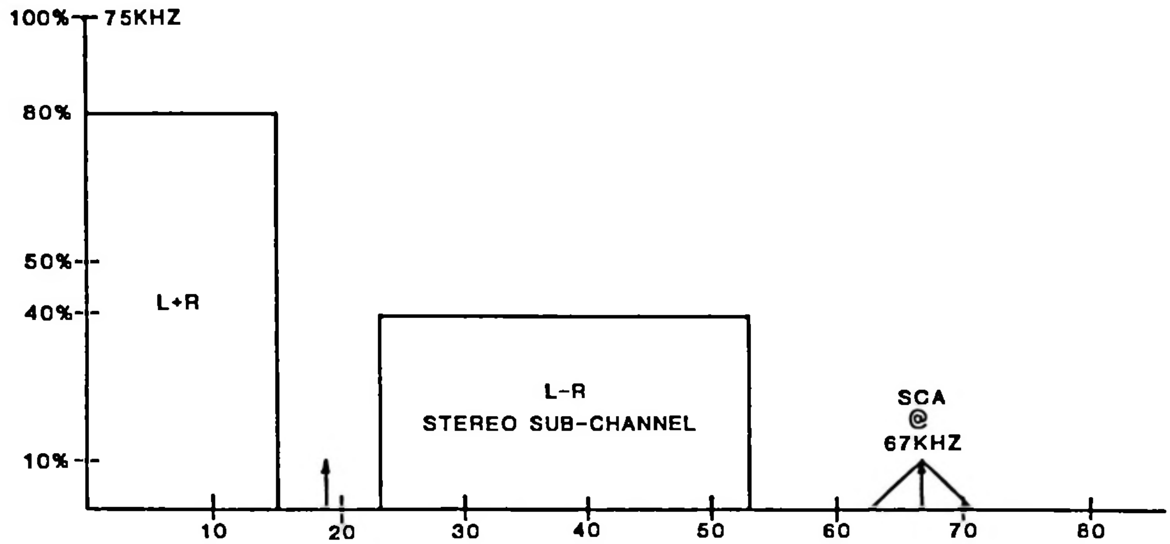
RESPONSE OF 2 POLE BUTTERWORTH FILTER

FIGURE A

```
FM ANALYSIS
HOUSTON MULTIPLEXER FILTER
DEMODULATED PILOT PHASE IS   -6.75 DEGREES
STEREOPHONIC CASE
ML= 100.0%  FL= 5. KHZ  MR=  0.0%  FR= 5. KHZ
INCIDENTAL AMPLITUDE MODULATION:   0.002 %   -88.41 DB (NOT DE-EMPHASIZED)

*****  DETECTED SPECTRA BEFORE DE-EMPHASIS  *****
FREQ  LEFT       DB        RIGHT      DB        L+R        DB        L-R        DB

 0   0.000084   -81.46    0.000085   -81.46    0.000000  -156.87    0.000084   -81.46
 1   0.000217   -73.29    0.000214   -73.39    0.000001  -117.81    0.000215   -73.34
 2   0.000009  -101.03    0.000009  -100.75    0.000000  -133.00    0.000009  -100.89
 3   0.000002  -114.92    0.000002  -114.38    0.000000  -138.03    0.000002  -114.46
 4   0.000013   -97.91    0.000018   -94.90    0.000000  -108.19    0.000015   -96.43
 5   0.999225    -0.01    0.001156   -58.05    0.499999    -6.02    0.499227    -6.03
 6   0.000055   -85.13    0.000050   -86.00    0.000003  -110.67    0.000053   -85.56
 7   0.000004  -107.68    0.000004  -107.58    0.000000  -130.80    0.000004  -107.65
 8   0.000009  -100.49    0.000009  -101.28    0.000000  -127.45    0.000009  -100.89
 9   0.000394   -68.09    0.000381   -68.29    0.000075   -82.50    0.000362   -68.83
10   0.000074   -82.63    0.000336   -69.53    0.000203   -73.84    0.000131   -77.67
11   0.000019   -94.47    0.000017   -95.40    0.000001  -119.97    0.000018   -94.92
12   0.000009  -101.36    0.000008  -101.98    0.000001  -124.45    0.000008  -101.68
13   0.000010  -100.05    0.000002  -114.50    0.000004  -107.84    0.000006  -104.48
14   0.000130   -77.75    0.000234   -72.60    0.000102   -79.85    0.000182   -75.80
15   0.000088   -81.51    0.000119   -78.50    0.000315   -70.02    0.000361   -68.84

*****  TOTAL DISTORTION TO 15 KHZ AFTER 75 MICROSECOND DE-EMPHASIS  *****
RIGHT CHANNEL VALUE IS ROOT SUM SQUARE OF ALL COMPONENTS TO 15 KHZ
    LEFT:        0.061 %  RIGHT:       0.052 %  L+R:        0.033 %  L-R:        0.110 %
STEREO SEPARATION:  57.59 DB
FUNDAMENTAL SEPARATION L TO R:  58.40 DB
DISTORTION IN R DUE TO L:  0.056 %

SCA SIGNAL EXTRACTED BY MEANS OF IDEAL 20 KHZ BANDWIDTH FILTER
SCA ANALYSIS
DEVIATION= 6.000 KHZ  FREQ= 8 KHZ
100% MODULATION DEFINED AS  6.00 KHZ DEVIATION
TRANSMITTED INJECTION=10.00 %
DETECTED INJECTION   = 9.92 %
INCIDENTAL AMPLITUDE MODULATION:   1.180 %   -38.47 DB (NOT DE-EMPHASIZED)

*****  DETECTED SPECTRUM BEFORE DE-EMPHASIS  *****
FREQ  SCA        DB

 0   0.000000  -200.41
 1   0.000238   -72.45
 2   0.000012   -98.51
 3   0.000056   -85.08
 4   0.065783   -23.78
 5   0.001920   -54.33
 6   0.002800   -51.06
 7   0.000092   -80.68
 8   0.000127   -77.95
 9   0.005848   -45.25
10   0.009727   -40.24

SCA UNMODULATED:
RECOVERED (CROSSTALK) MODULATION TO 10 KHZ AFTER 150 MICROSECOND DE-EMPHASIS:
       0.202 %   -53.88 DB
```

SAMPLE COMPUTER
ANALYSIS OUTPUT

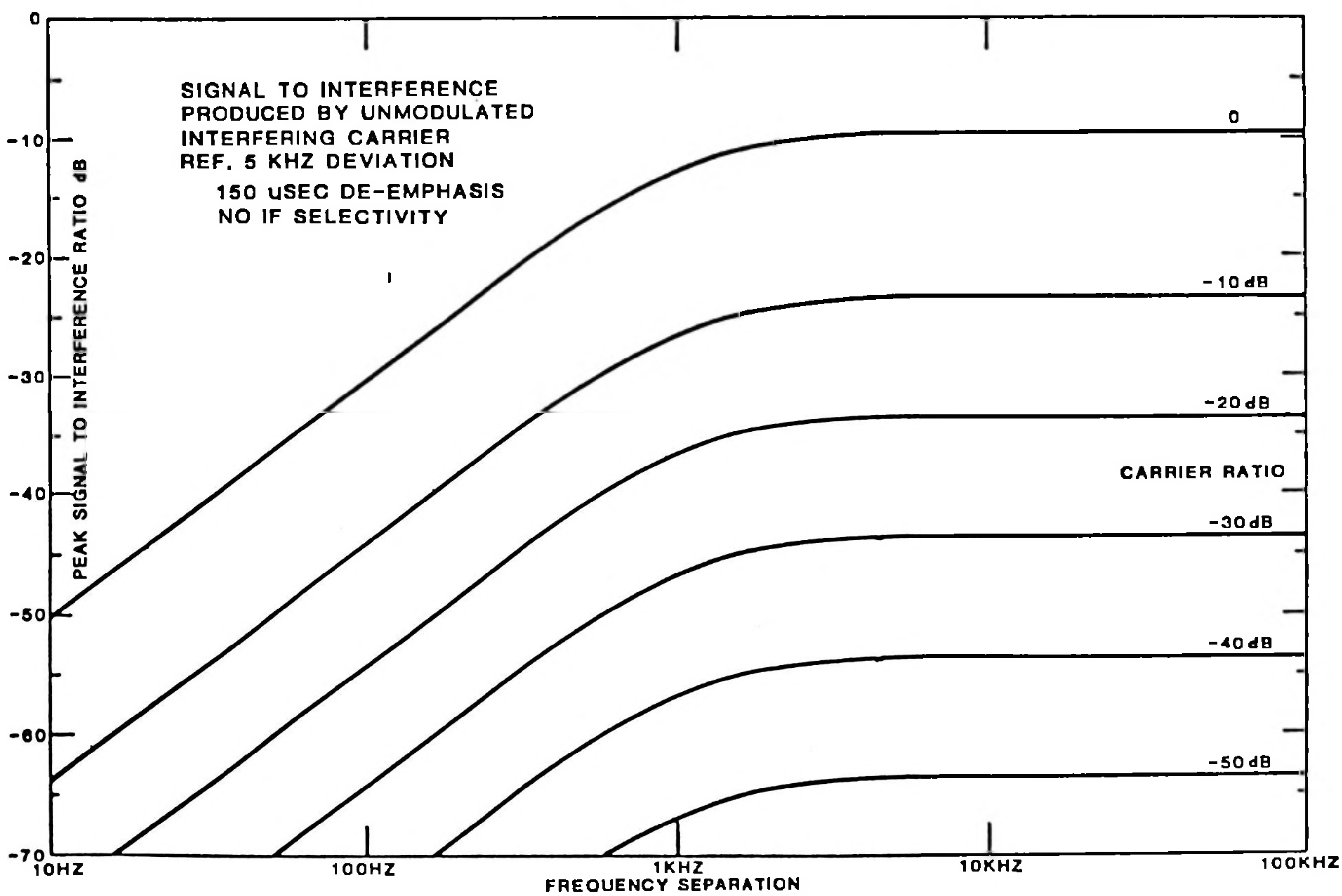
FIGURE B

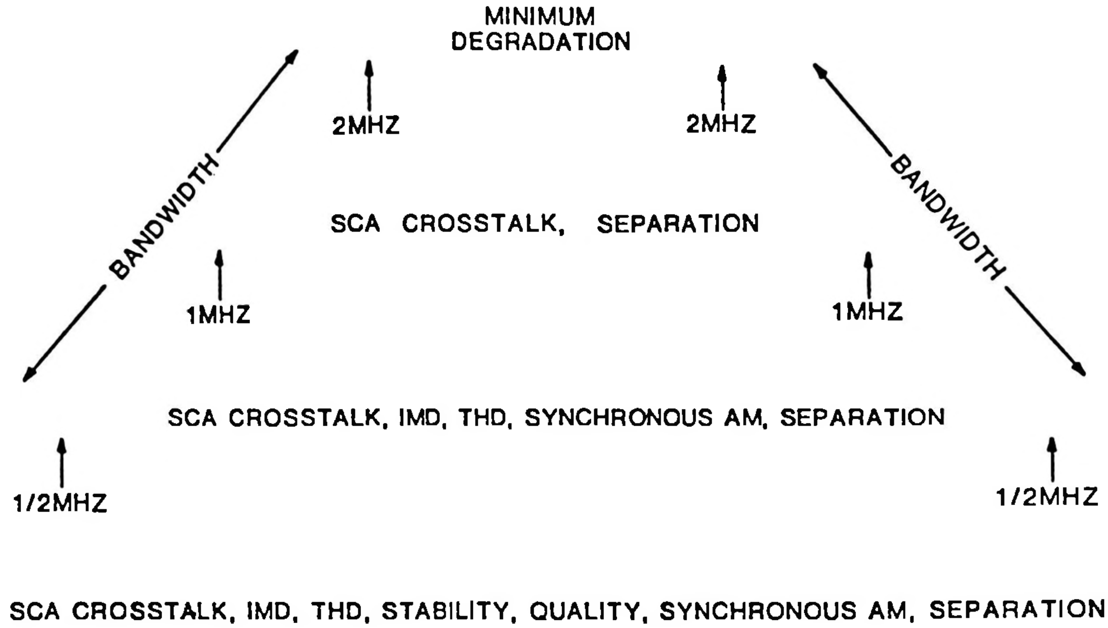


SPECTRUM OF COMPOSITE
FM STEREO BASEBAND

FIGURE C

FIGURE D

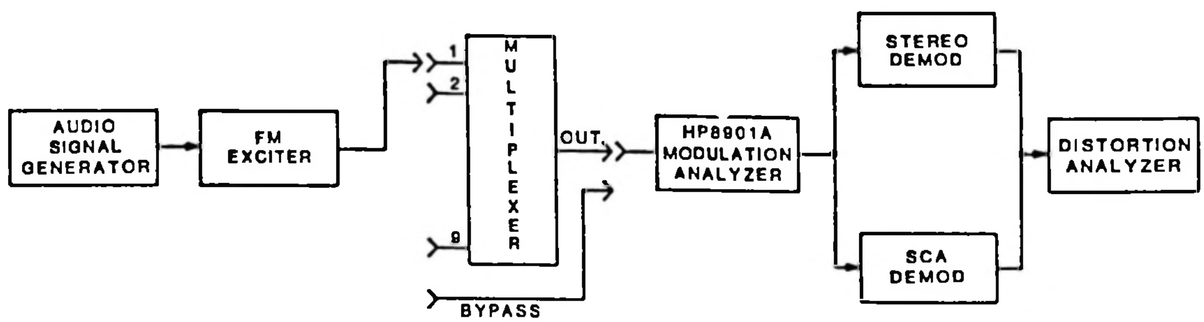




PROBLEM PYRAMID

FIGURE E

FIGURE F



TEST EQUIPMENT SET-UP
FOR MEASURING FM
PERFORMANCE DEGRADATION

MULTIPLEXER FM PERFORMANCE

FM performance requirement for a passive component? A multiplexer with 50dB stereo separation? Yes, this type of performance can be specified for a multiplexer. Furthermore, attention to FM performance will pay off in the field in delivery of quality signals.

To adequately power combine nine FM transmitters and assure premier FM performance, some additional design and test measures had to be taken. First, some stringent goals were set for the allowable degradation in FM performance. Then the multiplexer became an object of careful design, analysis, measurement, review, and modification. Because of the various trade offs possible, (isolation, VSWR, bandwidth, etc.), repeated design/measurements were often necessary.

The process of frequency modulation (FM) is inherently more complicated than that for amplitude modulation (AM), and the mathematical messiness of FM does not lend itself to easy analysis. Described simply, FM is the variation of an instantaneous carrier frequency in accordance with the information to be transmitted. The transmitted information is spread about the RF carrier in a multiplicity of sidebands whose amplitudes and phases are determined by Bessel functions. The FM signal theoretically occupies infinite bandwidth. In practice, however, truncation of the FM spectrum's insignificant terms is necessary in order to make the system workable. Performance of FM is determined by RF bandwidth. Both amplitude and phase (or group delay) of the spectral components will modify an FM signal. The various cavities found in a multiplexer provide ample circuitry to vary the amplitude and phase over the RF bandwidth of interest. Frequently, single reject cavities are used in multiplexers, but to achieve high isolation and broader RF bandwidths, a double tuned cavity was used. The typical response for both single and double tuned cavitites are shown in Figure A.

The FM performance analysis technique involved the use of a complex computer program and measured RF parameters. A set of measured multiplexer RF data of amplitude and group delay were used as input for the computer. The multiplexer input with 800kHz spacing was selected as a worst case design. The computer program was then run to determine FM performance characteristics such as THD, separation, crosstalk, etc. The bandwidth of the FM signal to the "significant" sidebands was examined. To assure conservative prediction accuracy, a calculation bandwidth of 0.5 MHz was used. The analysis, (sample shown in Figure B), revealed that stereo THD should be less than 0.1%, stereo separation of 50dB, and SCA crosstalk about 50dB. However, data and theory needed to be checked and correlated with observations and measurement.

INTERMODULATION PRODUCTS

The baseband signal of stereo FM with SCA is shown in Figure C. RF bandwidth limiting will generate products in the stereo SCA baseband in accordance to:

$f_{IM} = \pm mf_1 \pm nf_2 \pm$............. EQ.1

Where:

$f_1 f,$ . . . . . = Frequencies present in the baseband

n, m . . . . = integers 0 to ∞

f_{IM} . . . . . = IM product

n + m + ...= order of IM product

The number of IM products can be very high. By considering a Left, a Right and an SCA signal, 512 second and third order IM products are possible. Equation 1, however, does not predict the magnitude of the IM. RF bandwidth limiting creates these products but because of de-emphasis of the audio, (mono, stereo, and SCA), the prediction of the effects of these are complex. For example, the affect of a single IM product in the SCA band is shown in Figure D. As shown in Figure D, IM frequencies close (100 Hz) to 67 kHz produce little SCA degradation. But strong IM signals of 1 kHz or greater spacing from 67 kHz will degrade the SCA performance. According to Figure D, to maintain an SCA signal to interference level of 50dB, all IM levels must be >38dB down from the SCA injection or 58dB down from the peak multiplex signals. This 58dB (0.1%) requirement prescribes an excellent FM signal source and wide RF bandwidth.. The foregoing example demonstrates that the IM products generated by bandwidth limiting must be low.

PERFORMANCE INFLUENCES

To check the multiplexer system FM performance some important factors need to be recognized before giving high credibility to the absolute accuracy of any measured data:

1. The signal source must have excellent linearity in stereo and SCA operation.
2. The FM demodulator linearity of the baseband demodulator must be excellent.
3. The stereo and SCA demodulator must indicate accurate readings.
4. The equipment must measure performance suitable for broadcast standards.

It is impossible to measure the actual degradation of all performance parameters with today's equipment at all frequencies. Thus, it is advantageous to measure only the most sensitive parameters. One then can base the "quality" of a circuit's performance on as few measurements as possible. Ideally, by using a single FM performance measurement, the quality of an RF circuit could be tied quickly to its suitability to broadcast use. By testing, analysis, and field experience, we have confirmed that some parameters are better indicators of performance than others.

<u>We know that you cannot predict FM performance</u>:

1. By using RF bandwidth alone unless it is over 2 MHz bandwidth, (then the degradation is virtually beyond measurement).
2. By using group delay alone unless it is well behaved over a 2 MHz bandwidth, (same as No. 1).
3. By THD, response, mono to stereo sub channel crosstalk, SCA distortion, or S/N.

<u>You can predict FM performance:</u>

1. By using RF bandwidth **and** group delay.
2. By stereo IMD, stereo to SCA crosstalk, and perhaps, high frequency separation.

The quality of an RF circuit for FM can be most easily judged by the sensitive parameter of stereo to SCA crosstalk. A "problem pyramid" was constructed to indicate the progression of FM performance problems, (see Figure E).

MEASURED MULTIPLEXER DATA

The performance of the multiplexer was measured for mono, stereo, and SCA modes of operation. Baseline data was run to determine the lower limits of the test equipment by using a length of coax in place of the multiplexer (Figure F). No attempt was made to achieve extremely high performance and then measure the difference. The logic for this hinges on:

- The test equipment should be at least 10dB better than the parameters measured. Such equipment is not available.

- Complimentary modulators and demodulators can give artificial "good" readings.
- The sensitive parameters (IMD and stereo to SCA crosstalk) are most easily measured and indicate the ultimate performance of the circuit, thus negating the high performance requirement.

The source for the FM signal was a Harris MS-15 exciter, not particularly chosen or "hand tweaked". The wide band FM demodulator was a HP 8901A Modulation Analyzer. The HP 8901A output was then demodulated with special Harris FS-80/FC-80 monitors (stereo/SCA de-mods). Performance data then was taken for each transmitter input port to antenna output.

The measured mono, stereo, and SCA performance is shown in Table 1.

I. MONO PERFORMANCE

A.	THD @ 100 Hz	0.1%	
	@ 1 kHz	0.1%	
	@ 15 kHz	0.2%	
B.	IMD	0.1%	

II. STEREO PERFORMANCE

		THD	SEPARATION
A.			
	100 Hz	0.1%	> 50dB
	1 Hz	0.1%	> 50dB
	15 kHz	0.2%	∿ 50dB
B.	IMD	0.2%	

III. CROSSTALK

L + R→SCA	50dB
L + R→SCA	50dB
L / R→SCA	55dB

SRTG MULTIPLEXER FM PERFORMANCE
TYPCIAL DATA

TABLE 1

CONCLUSION

The relevance of the FM performance analysis and testing was that a maximum level of FM degradation was assigned to a piece of passive circuitry. With good FM transmitters, this technique assures minimum levels of FM performance before on-air-test. Understand that these measurements are **not** the same as sampling the transmitter before entering the multiplexer. These results were measured **through** the multiplexer. This correctly assigns the degradation to signal in the proper manner. By using a high performance FM exciter and transmitter, this designed-in high quality will be delivered to the broadcast listener with minimum affects attributed to the multiplexer.

FUTURE DESIGNS

Completion of this design has led to several significant concepts in analysis and hardware. First and significantly, we as industry are recognizing the system approach to equipment specifications. Even passive circuits can be given design limits normally attributed to sources. Secondly, stations can be frequency spaced as close as 800 kHz, achieve over 50dB isolation, and have excellent FM performance. Third, overall performance quality of an FM circuit can be determined by a one or two parameter measurement.

ACKNOWLEDGEMENT

We would like to express our appreciation to the following:

Dave Hershberger
Ed Westenhoder
Mike Nowak
Tony Uyttendael
Don Millar

REFERENCES

1. "Considerations in the selection of a 9-station FM combiner and antenna for SRTG", A. G. Uyttendaele, TAB; 1982.
2. "Amplitude bandwidth, phase bandwidth, incidental AM, and saturation characteristics of power tube cavity amplifiers for FM". R. Weirather and D. Hershberger, 1981 IEEE Broadcasting Conference.
3. "Handbook of Filter Synthesis", A. Zverev, John Willey and Sons.
4. "Intermodulation Products with Triplexed Solid State FM Transmitters", Harris Corporation, ABU Technical Review, May 1981.
5. "A Study of RF Intermodulation Between FM Broadcast Transmitters showing Filterplexed or Co-located Antenna Systems", G. Mendenhall, 1982 IEEE Broadcasting Conference.

USE OF SECTIONALIZED ANTENNAS FOR AM

By John R. Furr

Technical Director,
Clear Channel Communications, Inc.
Associate Engineer,
Diversified Communication Engineering, Inc.

San Antonio, Texas

USE OF SECTIONALIZED ANTENNAS FOR AM

The specific type of antenna used at any AM station is chosen to fit different needs:

1. The FCC establishes minimums for different classes.
2. The FAA establishes maximums for different sites.
3. The cost versus signal gain sets a practical limit.
4. Zone restrictions determine type and height.
5. Drive point impedances and stability are factors.
6. Coverage characteristics establish criteria.

Parts of these criteria for tower choice will be covered in this presentation. The majority will concentrate on the coverage characteristics. The specific needs will be initially addressed for WOAI in San Antonio, Texas, a Clear Channel station, then other applications will be shown.

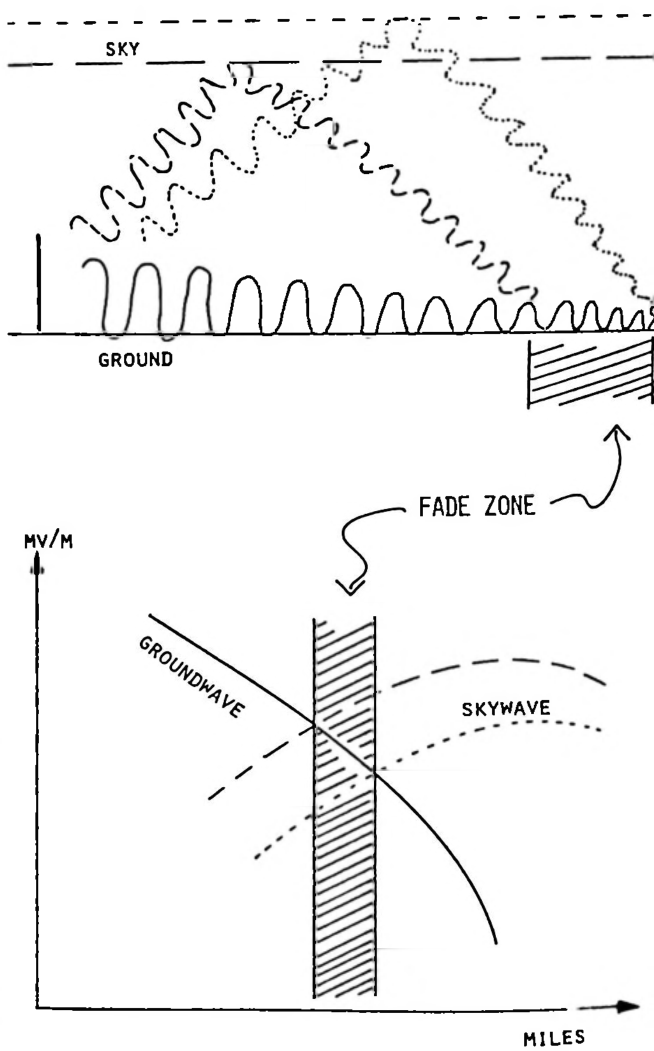
Clear Channel stations enjoy greater protection than other classes and are given the burden of maximumizing their facilities. It is to the licensee's and to the public's interest to maximize Day and Night coverages.

Dependable Daytime coverages can be expected to the 0.5 mV/m contour. Nighttime skywave signals run typically around 1 mV/m between 400 and 600 miles from the station. With changes in the atmosphere, this can rise to as much as 3 mV/m for short periods of time.

There is a point at which the ground wave ceases to predominate and the skywave is the strongest signal source. At the point where the skywave and the ground wave are equal in strengths, there is a cancellation of signals. You may have observed this same effect on the FM broadcast band where multipath signals arrive at equal strength at the receiver causing signal loss by cancellation. In the AM band, there is a whole zone, or ring, that is referred to as the distortion zone or fade zone. This is because the environment creating the skywave reflection is constantly changing.

Should you be listening to a station and the fade point pass your receiver, you would first hear one sideband cancelled, then the carrier, then the other sideband. If you listened with a communications type receiver

ILLUSTRATION OF FADE ZONE



with a beat frequency oscillator, you could turn this oscillator on and receive the station in what is called "exhalted carrier" mode. With this you could minimize the effect of the fade because this BFO is restoring the disappearing carrier. This same fade is experienced when driving through a deep null of a directional station.

Skywave phenomenon reacts differently in the nighttime than it does in the daytime. The standard broadcast band operates in the medium frequency range. The ultra-violet and other rays from the sun react with the upper layers of the atmosphere to cause these medium frequencies to be absorbed.

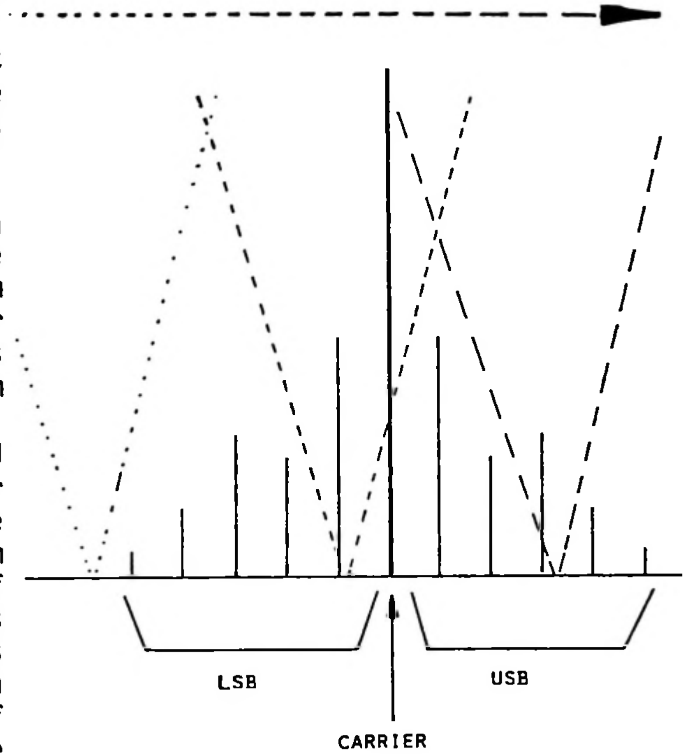
At night this absorbtion is reduced and some of the energy reaching the upper atmosphere is reflected to the earth. These atmospheric layers that reflect the medium wave frequencies are in a constant state of change. To quantify these changes, numerous tests were conducted in the 1920's and 1930's to establish some standards. Over a given period of time, there are two states of atmosphere that are of concern to the broadcasters: The state that exists fifty percent of the time and the state that exists ten percent of the time. From these designations, there have been established graphs in the FCC Rules to predict skywave coverages.

Before skywave coverage can be predicted, the vertical radiation characteristics of a given antenna must be known. The signal radiated from an antenna is determined by the current distribution. This radiation forms a pattern much like a donut sliced lengthwise and laid flat on its cut side.

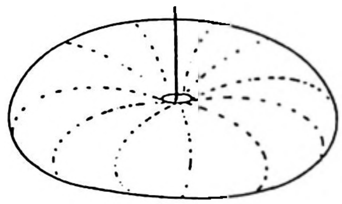
The skywave at any given point "P" located at a distance "D" from the antenna is calculated on the basis of the inverse signal radiated at the pertinent vertical angle, "theta". There are three vertical angles of interest: the 50% angle, the 10% high angle, and the 10% low angle. When the inverse field is known at each of these vertical angles, then the skywave is easily predicted.

From the 1930's until the late 1950's, WOAI used a half-wave antenna as did most all other Clear Channel stations in the USA. This is the vertical pattern of a 185 degree antenna. Think of this pattern as being a pie-shaped slice out of the donut around the

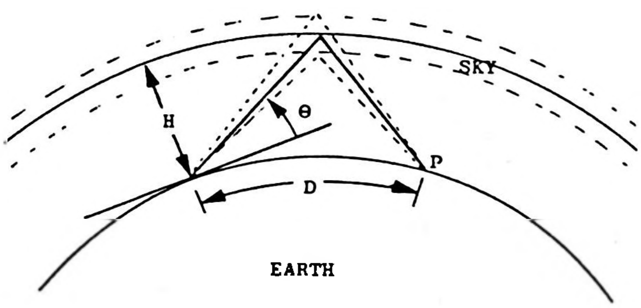
PASSING OF FADE WALL



VERTICAL "DONUT" PATTERN



PERTINENT VERTICAL ANGLES



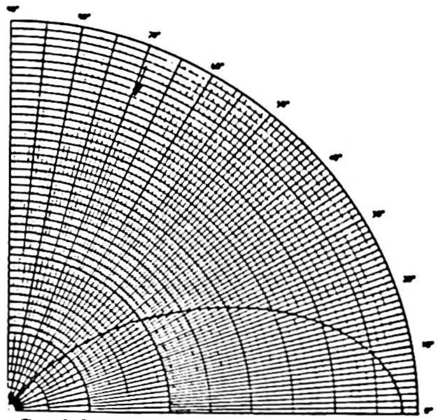
H = Height of atmospheric layers which reflect skywaves. These heights vary for 50% and 10% of the time.

D = Distance from transmitter site to point "P"

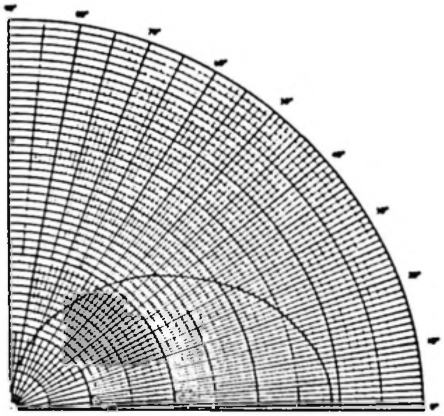
θ = Pertinent Vertical Angle for distance "D". Signal at this angle determines level of skywave at point "P" tower as viewed from the side. The local listeners are receiving the signal along the ground, or zero degrees vertical.

If we could go up in the air with a field strength meter over perfectly conducting earth and graph the readings at one mile, we would have the graphs shown on the right. Compare it to the commonly used quarter-wave antenna, also shown. Notice for the same amount of power, 1000 watts, put into the tower, less signal is radiated along the ground, or zero degree vertical, (194.9 mV/m) than the half wave (240.6 mV/m). We can now think of the vertical pattern changing as we change the antenna, as if it were a gelatin material placed in a plastic bag. The volume of gelatin (1000 watts) does not change, only the shape. As we reduced the height of the tower, the signal is pulled in along the ground and pushed more up into the sky. Compare it again to the slightly over half-wave antenna. Notice a very small second lobe is starting to form. It is made by the null located near 70 degrees vertical. (Notice that the nulls will be marked in all illustrations with an arrow.) We will examine more of these secondary vertical lobes in this study.

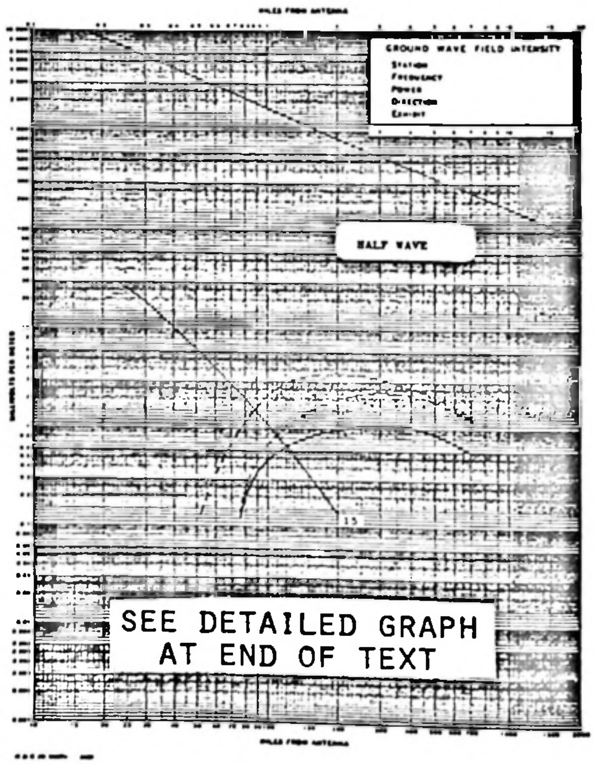
The conductivity graph shown here is with the half-wave antenna we just examined operating at 50,000 watts. Most of the populated area around San Antonio is located in soils that have a conductivity of 15 mmho. Because of our initial criteria, we will examine this soil area first. The inverse line is showing crossing the one mile line at a level we calculated from the 1000 watt pattern (240.6 mV/m) operating at 50,000 watts. This inverse is 1700 mV/m. The smooth curve descending from left to right is the ground wave curve. The upward moving curve on the right is the skywave curve. The 50% skywave curve is shown with a solid line and the two 10% curves are shown with dashed lines. Based on this graph, we can predict that 50% of the time, the fade point will be located at 120 miles from the transmitter site. Ten percent of the time it will move to 110 miles or to 88 miles before it returns. The ground wave curve shows that the 0.5 mV/m ground wave is located at 130 miles from the transmitter. Radio listeners located between 88 and 130 miles would be better served by a stable ground wave only, rather than suffer with annoying fades.



G=185
Inverse=240.6 mV/m



G= 90
Inverse=194.9 mV/m



Therefore, based on the criteria of the soil and the location of listeners around San Antonio, the half-wave antenna is unacceptable.

The chief operator of WOAI during the 1950's was the late Charles Jeffers. Jeffers, in cooperation with Andy Ring and others, did extensive research on ways to improve ground wave service and maximize skywave service. Jeffers wrote a paper for the "Proceedings of the I.R.E.", November, 1948. In this paper he adds yet another criteria:

"The ideal antenna for a high powered broadcast station, providing both a ground wave and a skywave service, should have a vertical radiation characteristic such that the skywave signal does not interfere with the desirable ground wave service.....The Nighttime primary service area would then be as large as the Daytime area, since the fading or distortion wall would not be the limiting factor. It is further desirable that the skywave signal strength rise rapidly in order to limit the intense fading area to a narrow band around the station." The antenna developed from this research is now in use at WOAI.

In the late 1950's WOAI was both AM and TV. When a new tall tower site was built for TV, the AM transmitter was co-located so that one operator could cover both transmitters. F.A.A. restrictions dictated the new site for the 1500 foot TV tower. Unfortunately this site was located in a area of poor soil conductivity. WOAI Radio, now no longer commonly owned with TV, decided to move the site to improve the ground wave coverage. The new site is in more desirable clay-type soil that field strength tests show has excellent conductivity. A study which sought to re-think the Jeffers approach using improved formulas at computer speed was conducted to see if there were aspects of the original design that could be improved with this site move. This paper was developed from that study.

Multi-elements horizontally positioned are commonly used in directional antennas to control coverages in the horizontal and vertical planes. This is accomplished by setting nulls on either side of a line of towers. Position of the nulls and their depths act as controlling elements to shape the lobes.

I.R.E. ARTICLE QUOTED IN TEXT


PROCEEDINGS OF THE I.R.E.—Waves and Electrons Section

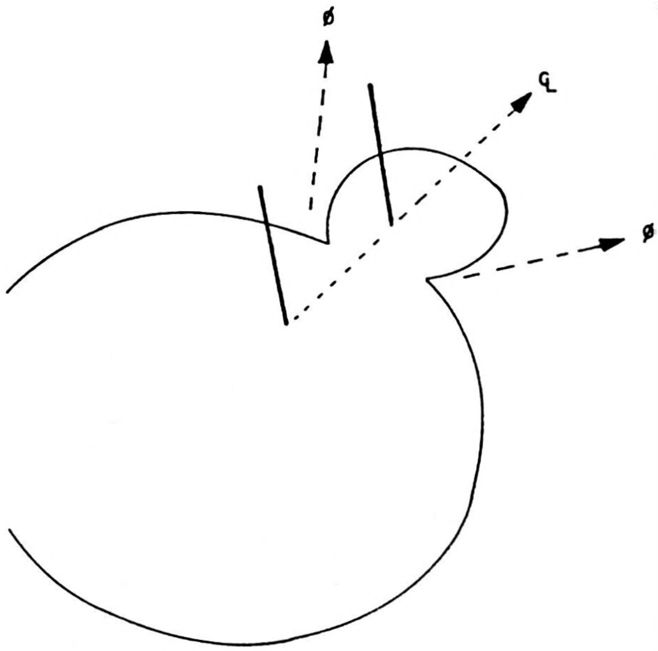
An Antenna for Controlling the Nonfading Range of Broadcasting Stations*

CHARLES L. JEFFERS†

INTRODUCTION

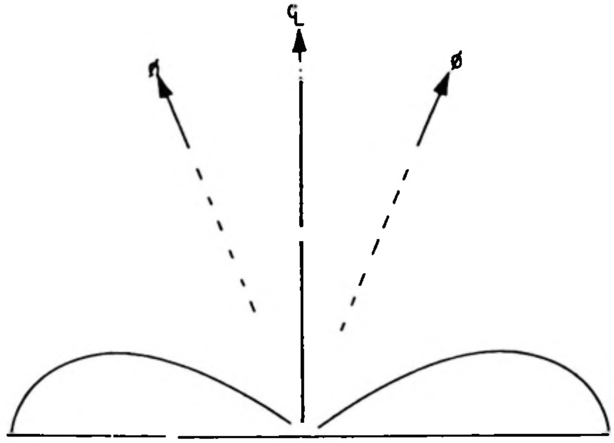
ELECTRICAL CHARACTERISTICS

HORIZONTAL PATTERN AND NULLS



With the sectionalized antenna, the same procedure is done, only the line of towers is stacked vertically and the null is adjusted only in the vertical plane. This adjustment is used to control the shape of the main lobe and keep it next to the ground.

For this study, we have used the F.C.C. formula to mathematically express the sectionalized antenna. This is the complete formula, including the hemispherical integration to determine the inverse at one mile at one kilowatt:



VERTICAL PATTERN AND NULLS

$$152.15158/\left(\left(\tfrac{1}{2}f(\theta)_0 + \sum_{5}^{85} f(\theta)^2 \cos\theta\right)(5\,/180)\right)^{\frac{1}{2}}$$

Where

$$f(\theta) = \frac{\sin W\,(\cos B \cos(A \sin\theta) - \cos G) + \sin B\,(\cos D \cos(C \sin\theta) - \sin\theta \sin D \sin(C \sin\theta) - \cos W \cos(A \sin\theta))}{\cos\theta\,(\sin W\,(\cos B - \cos G) + \sin B\,(\cos D - \cos W))}$$

Where

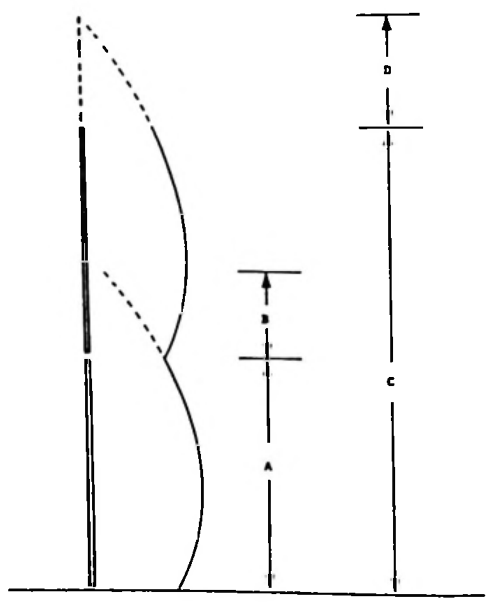
θ = Vertical Angle
A = Electrical height of lower tower section
B = Apparent electrical height (based on current distribution) of A
C = Electrical height of entire tower
D = Apparent electrical top loading (based on current distribution)
G = A + B
H = C + D
W = H - A

This is the F.C.C.'s definition of terms. Section "A" is the electrical height of the lower section. "B" is the apparent electrical suppression of current in the lower section. "C" is the overall height, or the sum of the upper and lower section. "D" is the electrical top loading.

Apparent suppression of the lower section can be explained, for simplicity's sake, as base loading. This is accomplished by placing tuned elements between the base and the ground system. The relative indication of "B" can be measured with an antenna monitor's indication of current ratios. The Jeffers approach to the design was to treat them as field ratios. The F.C.C. formula, however, treats then as apparent electrical suppression, in degrees.

The heading will show the parameters at which the calculations were made. Notice the field strength at one mile. As the sections are increased, much of the gain will come from

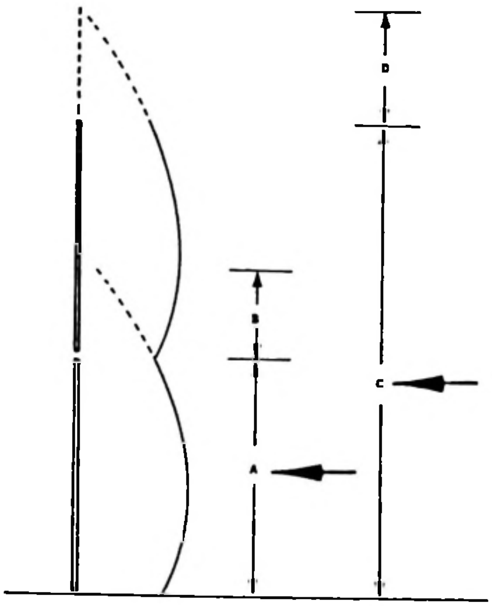
DEFINITION OF VARIABLES



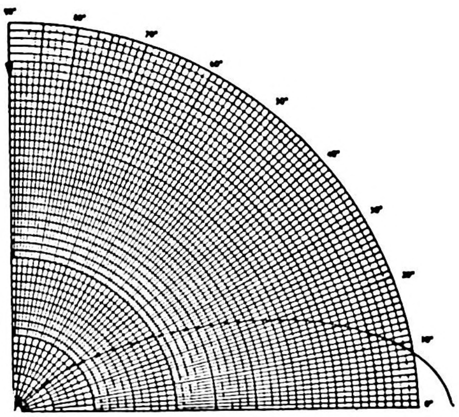
the natural suppression and improved efficiency that increased antenna height will provide from the form factor of a single element. Now we will examine the resulting vertical patterns produced when the sections of the antenna are changed in the increments shown. NOTE: some of the apparent suppressions shown in these examples may not be practical in the field, but are included here to get the larger perspective of the effect of current suppressions, or loading.

The first series of vertical patterns will show the effect of increasing both upper and lower sections of the tower, in equal amounts. "B" has been determined as five degrees, because it is reasonable to assume that the currents in the tower would not be maintained at perfectly equal levels in the field.

Watch the signal gain along the ground or zero degree angle. The arrows will indicate the null placement on the graph.

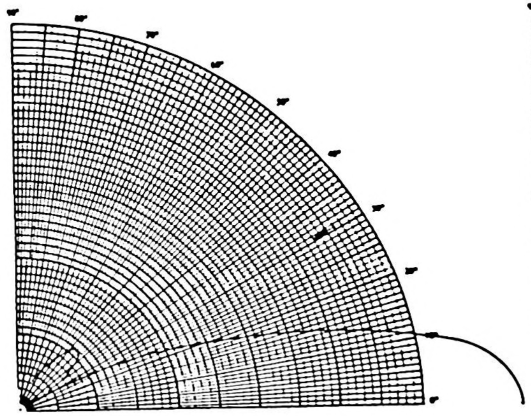


CHANGING A AND C



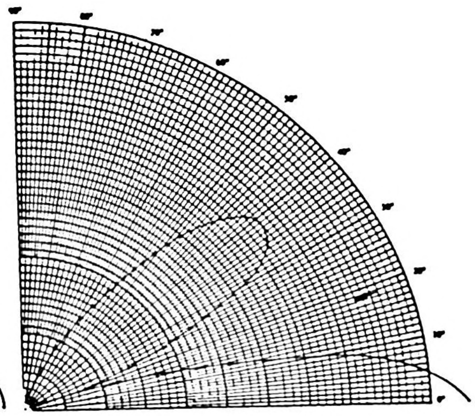
A=165
B= 5
C=330
D= 0

Inverse=
270.9 mV/m



A=175
B= 5
C=350
D= 0

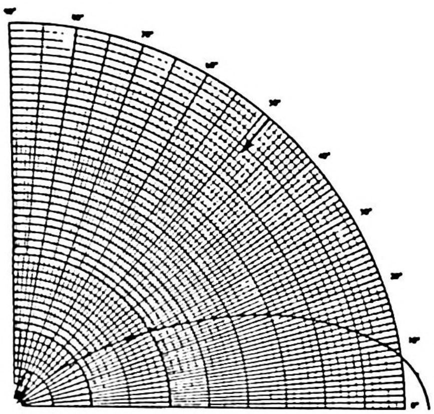
Inverse=
313.7 mV/m



A=180
B= 5
C=360
D= 0

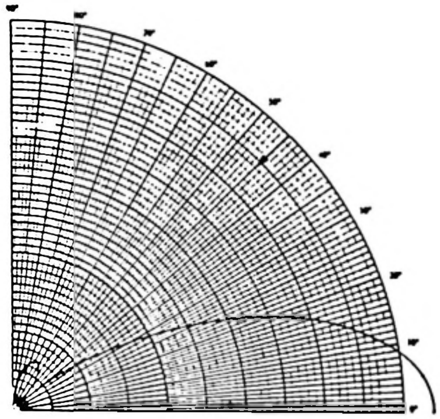
Inverse=
272.4 mV/m

This series of graphs display the vertical patterns with no change in A or C; A=120 and C=240. We will only change B, the apparant suppression of the lower section, and see how it controls the vertical null.



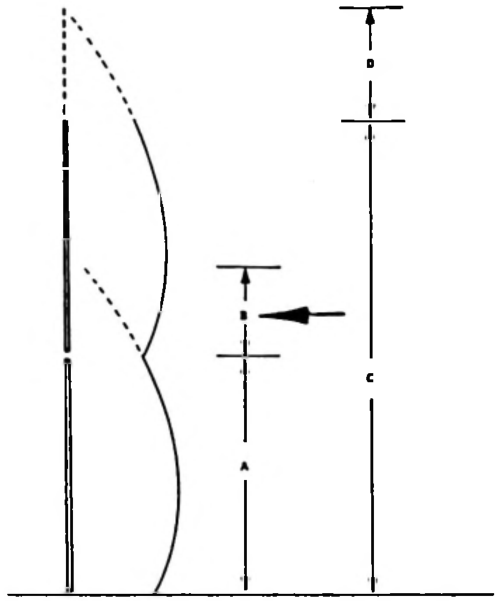
A=120
B= 60
C=240
D= 0

Inverse=
265.3 mV/m

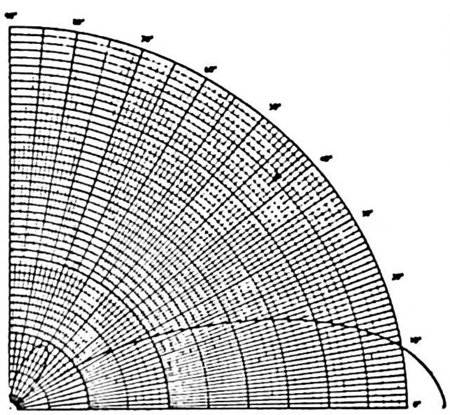


A=120
B= 70
C=240
D= 0

Inverse=
269.9 mV/m

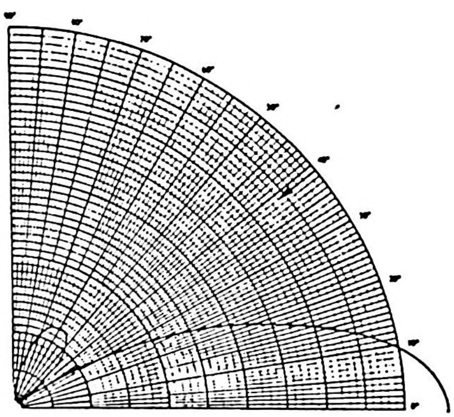


CHANGING B



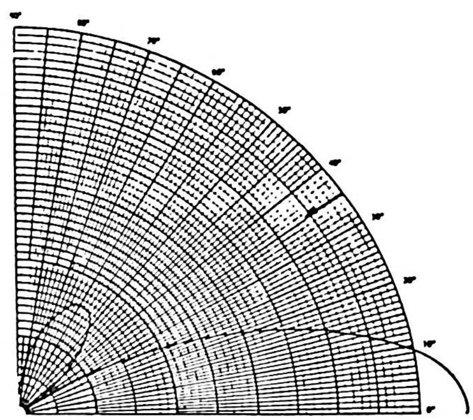
A=120
B= 80
C=240
D= 0

Inverse=
273.8 mV/m



A=120
B= 90
C=240
D= 0

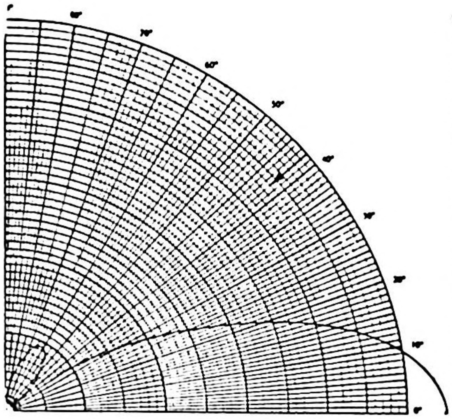
Inverse=
277.0 mV/m



A=120
B=100
C=240
D= 0

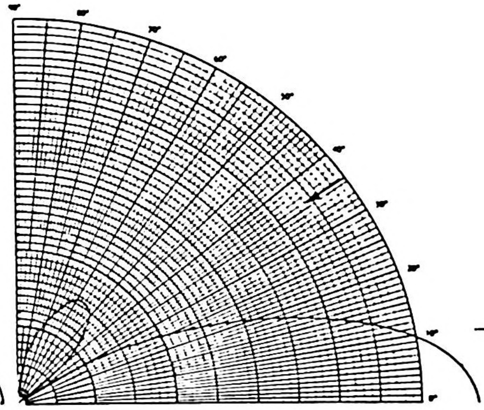
Inverse=
278.9 mV/m

This series of graphs will maintain B=80 degrees and change sections A and C in the shown increments. Look for a lot of horizontal gain and similar, but more rapid development of the upper vertical lobe.



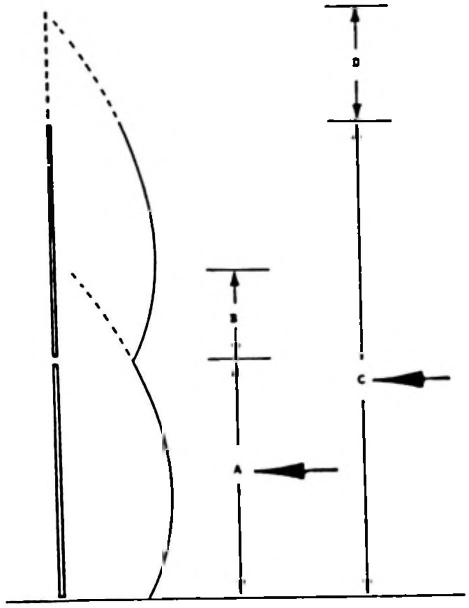
A=120
B= 80
C=240
D= 0

Inverse=
273.8 mV/m

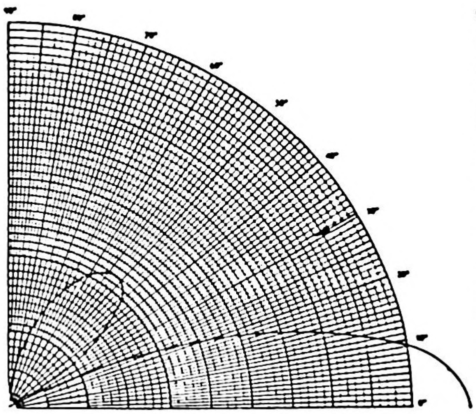


A=130
B= 80
C=260
D= 0

Inverse=
283.7 mV/m

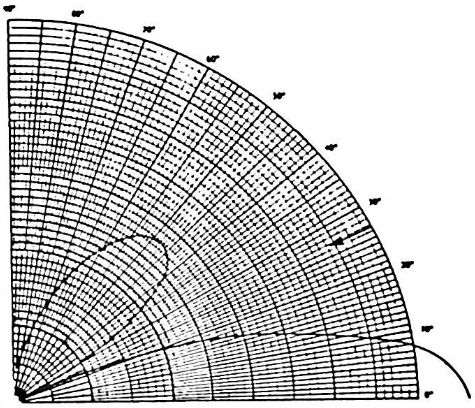


CHANGING A AND C



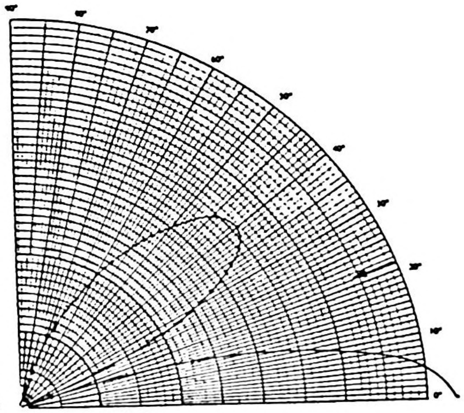
A=140
B= 80
C=280
D= 0

Inverse=
286.8 mV/m



A=150
B= 80
C=300
D= 0

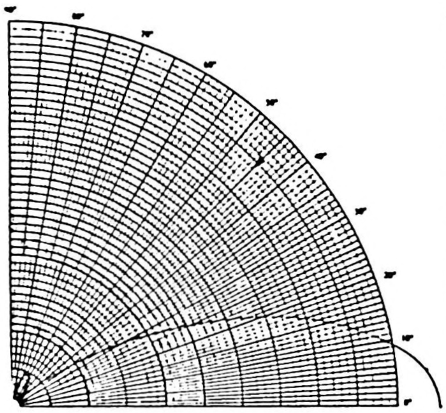
Inverse=
282.7 mV/m

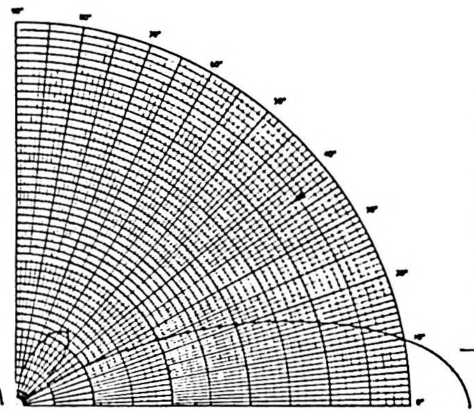


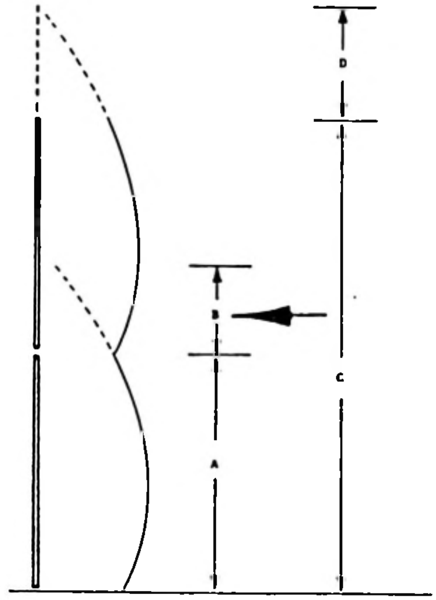
A=170
B= 80
C=340
D= 0

Inverse=
269.8 mV/m

This series of graphs seeks to maximize the horizontal gain along the ground, by raising the upper and lower sections to 140 degrees each (A=140, C=280). The apparent suppression, B, will be increased so we can study the change in efficiency.







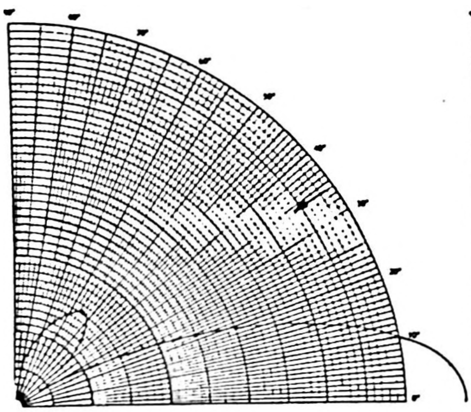
CHANGING B

A=140
B= 30
C=280
D= 0

Inverse=
275.5 mV/m

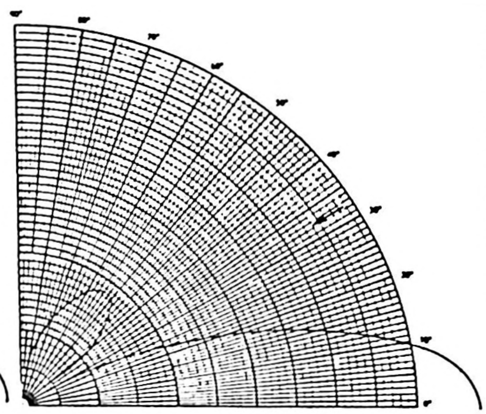
A=140
B= 50
C=280
D= 0

Inverse=
287.3 mV/m



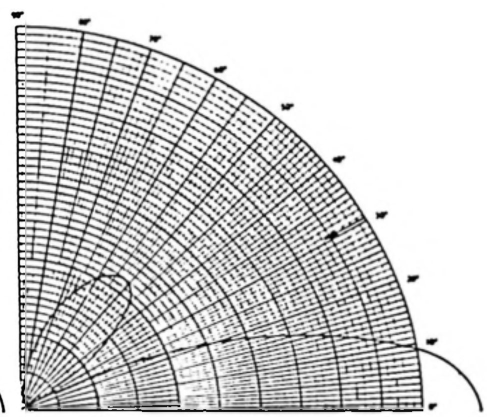
A=140
B= 60
C=280
D= 0

Inverse=
289.3 mV/m



A=140
B= 70
C=280
D= 0

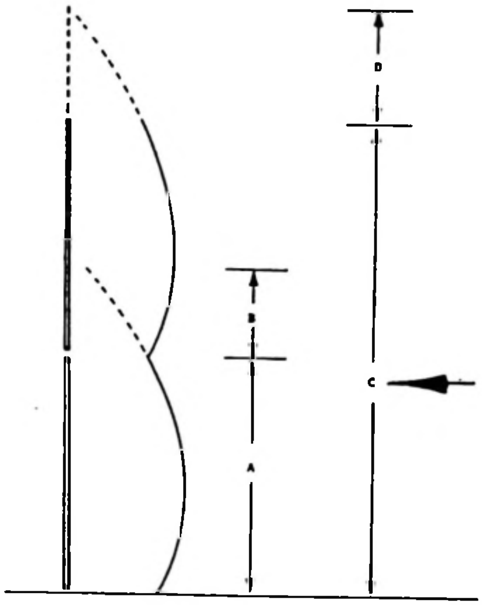
Inverse=
289.1 mV/m



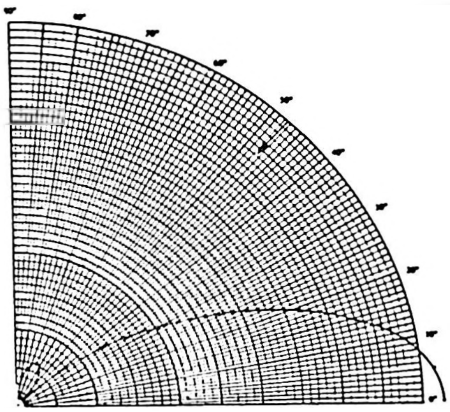
A=140
B= 80
C=280
D= 0

Inverse=
286.8 mV/m

It is also intersting to see what happens when one section is maintained and the other changed. In this case we will hold section A at 120 degrees, B will stay at 80 degrees, and the upper section increased, thus increasing C. (Although no graphs were prepared for top loading, the effect of increasing the upper section produces a pattern very similiar to the effect of increasing top loading, D, only.)

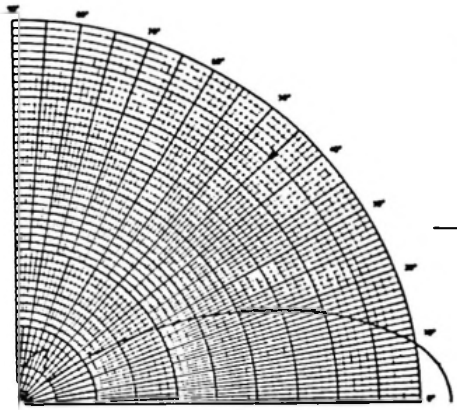


CHANGING C



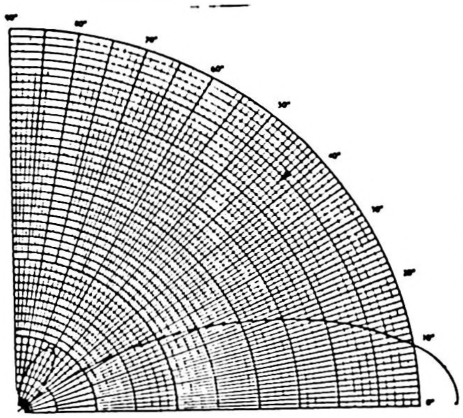
A=120
B= 80
C=220
D= 0

Inverse=
263.9 mV/m



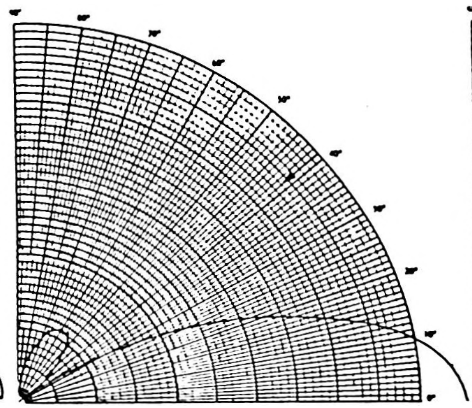
A=120
B= 80
C=230
D= 0

Inverse=
268.8 mV/m



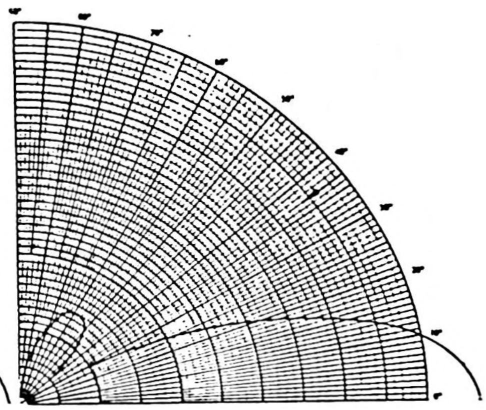
A=120
B= 80
C=240
D= 0

Inverse=
273.8 mV/m



A=120
B= 80
C=250
D= 0

Inverse=
278.9 mV/m



A=120
B= 80
C=260
D= 0

Inverse=
283.6 mV/m

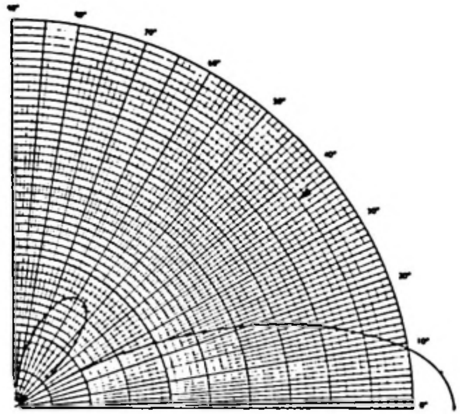
You have examined some examples of the various combinations that a sectionalized antenna can take in the vertical plane; which controls the horizontal efficiency. This is a pattern of the 225 degree, or the 5/8ths wave antenna. When an antenna height is increased past 180 degrees, as seen on this page, a null is formed naturally off the end of the tower. As the tower is increased more in height, the null moves down toward Theta=0, or ground, which creats a second lobe. This lobe is shown in the 5/8ths wave antenna.

If the antenna were increased beyond 225 degrees, the null would continue to move downward causing the secondary lobe to develop rapidly and quickly deteriorate the ground lobe to nothing.

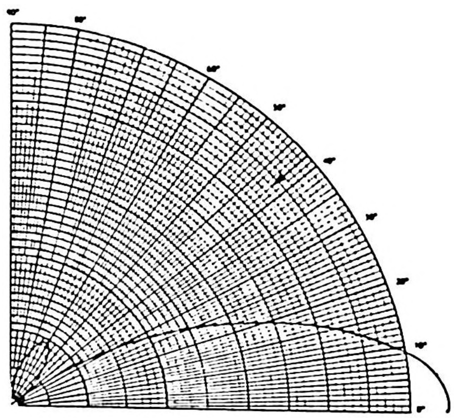
The inverse of the 5/8 wave antenna is 275.7 mV/m at one mile. Compare this to the sectionalized antenna with 175 degree sections and 5 degree suppression with an inverse of 313.7 mV/m. The sectionalized has a 38 mV/m inprovement at 1000 watts.

When nighttime skywave is considered, the two antennas have a contrasting prespective. Because the vertical null of a non-sectionalized antenna is dependent on the physical height of the tower, there is no control over the null placement. The sectionalized antenna can control the null placement and therefore adjust the size of the secondary lobe while achieving similar efficiency.

At the end of this text there are coverage graphs which show the groundwave at various conductivities. The skywave is overlayed so the extent and size of the fade zones can be readily seen. The graph for the sectionalized was calculated with A=120, B=80, C=240, and D=0. It has an efficiency at 1000 watts of 273.8 mV/m at one mile. These values were chosen to closely approximate the 5/8 wave antenna for comparison. With other adjustments, the vertical lobe on the sectionalized could be reduced. The half-wave is included to compare the vertical pattern verses the coverage graphs.

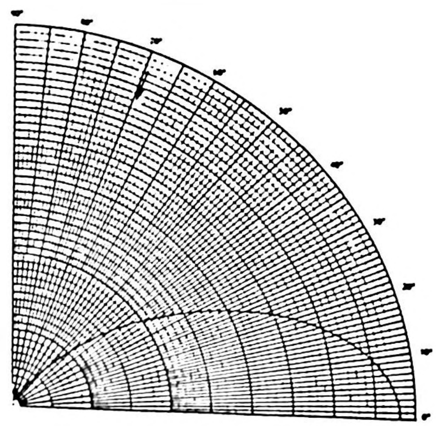


G=225
Inverse=275.7 mV/m



A=120
B= 80
C=240
D= 0

Inverse=
273.8 mV/m



G=183

Use the three graphs at the end of this paper to compare the three antennas for fade zone control. Jeffers stated that the skywave should build fast to restrict the outer limits of the fade zone. Compare the 50 per-cent curves (solid lines). Only the sectional antenna has a steep build which results from the near blunt vertical pattern within the first five degrees. The 5/8ths wave and half-wave do not have this rapid build.

Compare the half-wave to the sectionalized. There are intersections with the 15 mmho contour from 88 to 120 miles which is within the 0.5 mV/m ground contour. The sectionalized 50 per-cent curve is at 160 miles, 40 miles beyond the 0.5 mV/m ground contour. With this particular design, there is intrusion within the 0.5 mV/m ground wave 10 per-cent of the time but only comes as close as 92 miles. Other adjustments to this design could significantly reduce this intrusion.

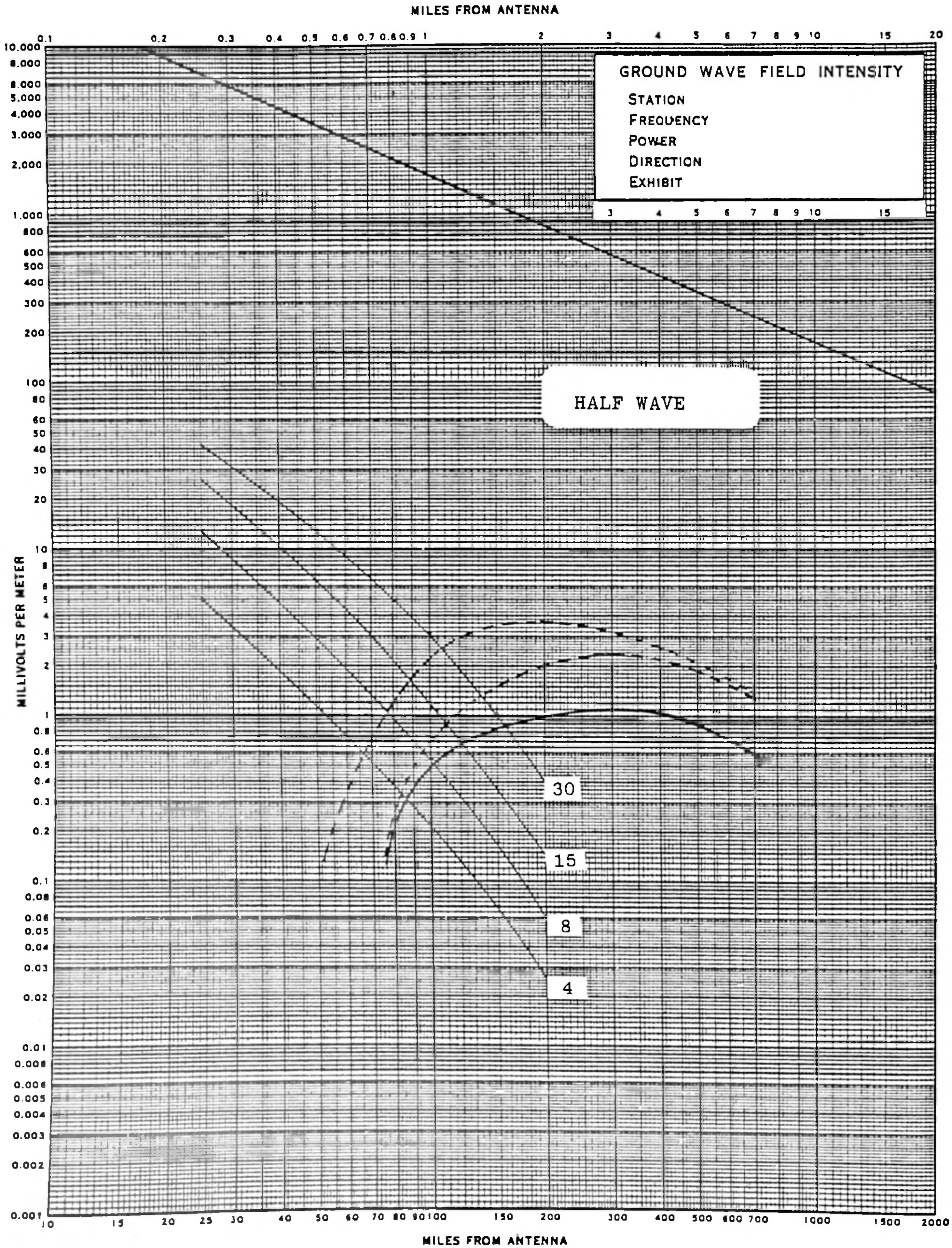
Examine now the 5/8th wave antenna. With soil areas that have less than ideal conductance, the skywave (50 %) is so close that there would be a constant fade problem from 110 to 175 miles. Ten per-cent of the time it would come to 78 miles. The fade zone with this antenna would be much more severe than would be with either of the other two antennas.

Regional stations are not afforded protection beyond the 2.5 mV/m nighttime contour. Nighttime skywave interference at the maximum 5000 watt power levels can be expected beyond this night service area. However, if the effective radiated power in the lobes become significantly large, fading could be experienced using the 5/8th wave antenna in areas of high conductivity. The coverage graph shows 10 per-cent signal rising as high as 3.1 mV/m at 105 miles.

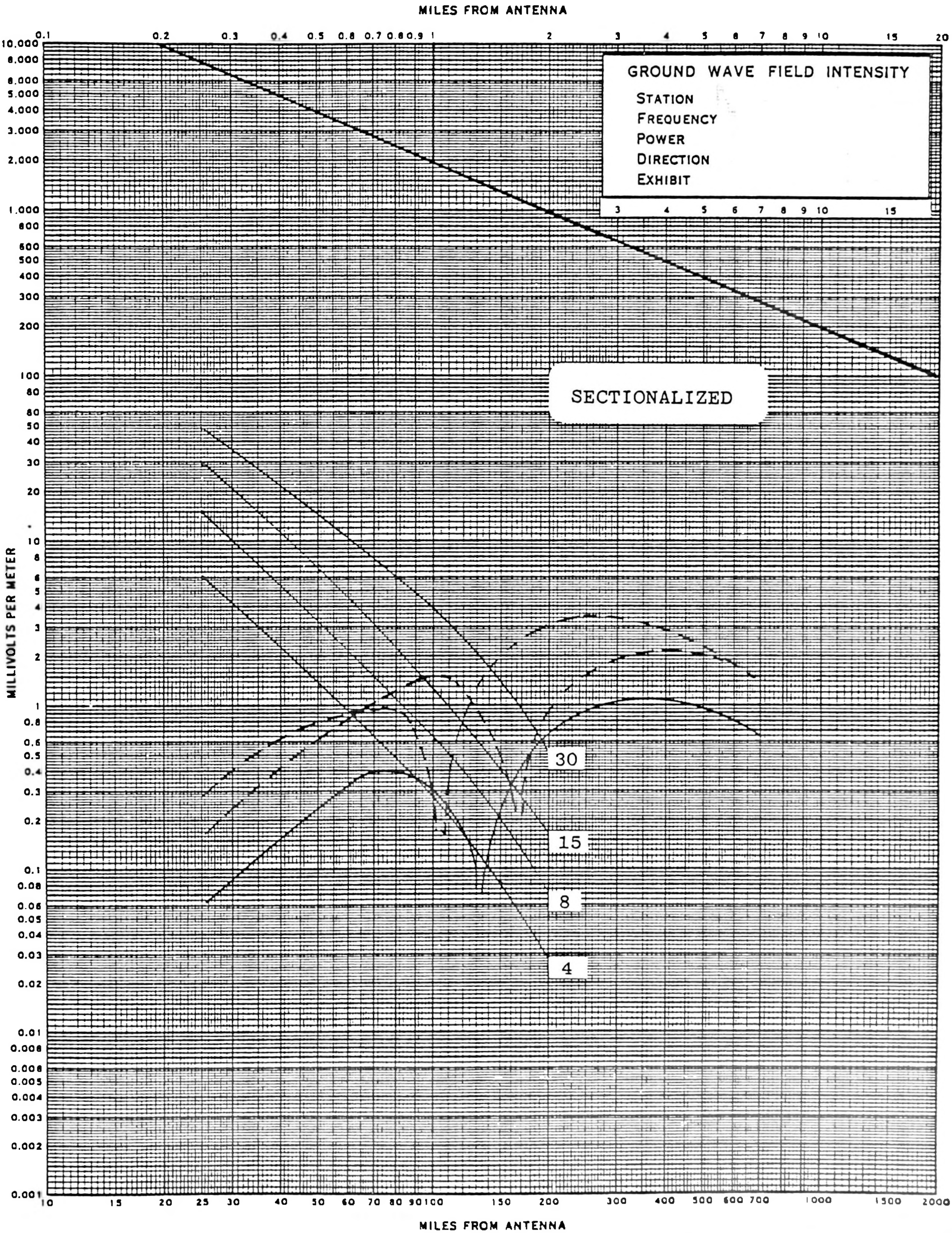
In designing nighttime directionals, there are often stations close by that need protection from high vertical angles of a directional. Antenna systems of four or more towers (sometimes three) can expect to have high levels of power radiated in the vertical plane where the ground level has low signal. This is especially true if near quarter wave antennas are used. Because of the vertical control the sectionalized antenna exhibits, it then becomes an option to meeting protections in cases where the pertinent angle is high and it is desired to increase the ground wave in the null areas.

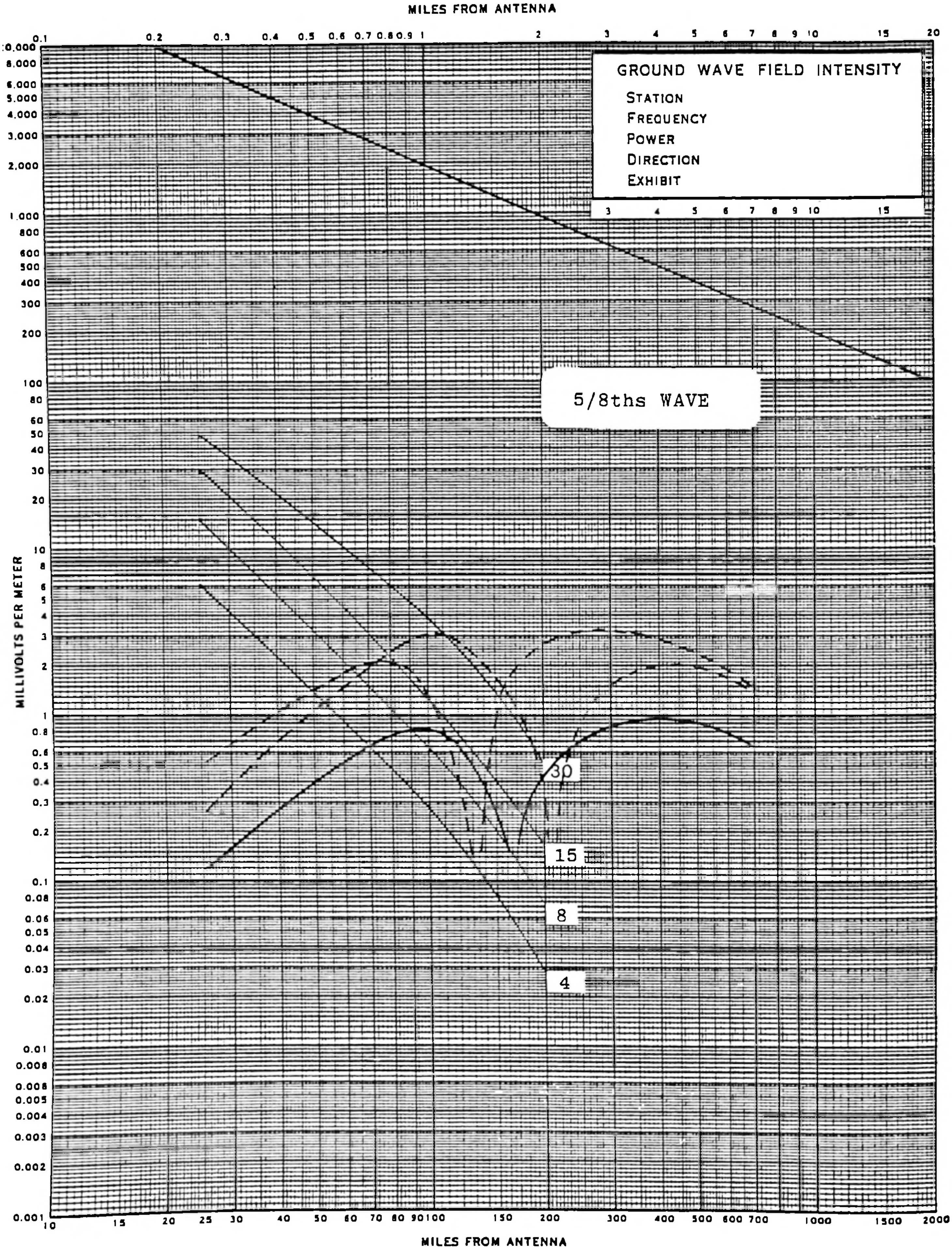
With today's problem of land availability in metropolitan areas coupled with environmental considerations, the sectionalized antenna offers another option to help extend service area with existing sites. If this alternative will work, the additional cost of steel and tuning networks can often be much less than land costs associated with relocation.

As with any antenna system, a design of a sectional must take into consideration the potential drive points, voltages, and currents. Physical separation of the elements with insulators which have the structural integrity and dialectric properties must be planned. Final operating stability should weigh heavily in the decision of element size.



MILES FROM ANTENNA
GROUND WAVE FIELD INTENSITY
STATION
FREQUENCY
POWER
DIRECTION
EXHIBIT
HALF WAVE
MILLIVOLTS PER METER
30
15
8
4
MILES FROM ANTENNA


MILES FROM ANTENNA
GROUND WAVE FIELD INTENSITY
STATION
FREQUENCY
POWER
DIRECTION
EXHIBIT
SECTIONALIZED
MILLIVOLTS PER METER
30
15
8
4
MILES FROM ANTENNA


MILES FROM ANTENNA
GROUND WAVE FIELD INTENSITY
STATION
FREQUENCY
POWER
DIRECTION
EXHIBIT
5/8ths WAVE
MILLIVOLTS PER METER
30
15
8
4
MILES FROM ANTENNA

SOLID-STATE TRANSMITTER TECHNIQUES DEVELOPED FOR CRITICAL AEROBEACON APPLICATIONS MAY BE USED FOR AM BROADCASTING WITH SIGNIFICANT ADVANTAGES

Dennis H. Covill

NAUTEL

Hackett's Cove, Nova Scotia

INTRODUCTION

The AM Broadcast band occupies the spectrum from 535 to 1605 kHz. Its lower spectral neighbour is the Aeronautical/Marine Navigation band which is allocated from 190 to 535 kHz, except for the Maritime Mobile spectral slot. Transmitters operating in this Navigation band generate a carrier signal with keyed tone modulation (for identification) and, in most cases, subsidiary voice modulation (for weather or other data). These transmitters operate at carrier power levels of up to approximately 5 kW depending upon the desired range, terrain, and local noise environment and, for obvious reasons, are referred to as Radiobeacons because of the principal purpose of the radiated carrier.

The loss of signal from a Radiobeacon could be a very serious matter, particularly in remote regions where it is the only radio navigation aid. For this reason, it had always been standard practice to duplicate the Radiobeacon in critical installations and to provide automatic failure sensing and changeover to a duplicate (standby) system when required. This was prior to the advent of wholly solid-state transmitters and the attendant "soft failure" technology. Another special problem for these transmitters arises from their frequent location in remote, difficult-access areas. Not only must they operate unattended with infrequent service visits, but the need to regularly transport fuel for the associated electrical generators poses logistic and economic difficulties. From these problems it may be reasonably deduced that Radiobeacons have a special need to be both super reliable and super efficient.

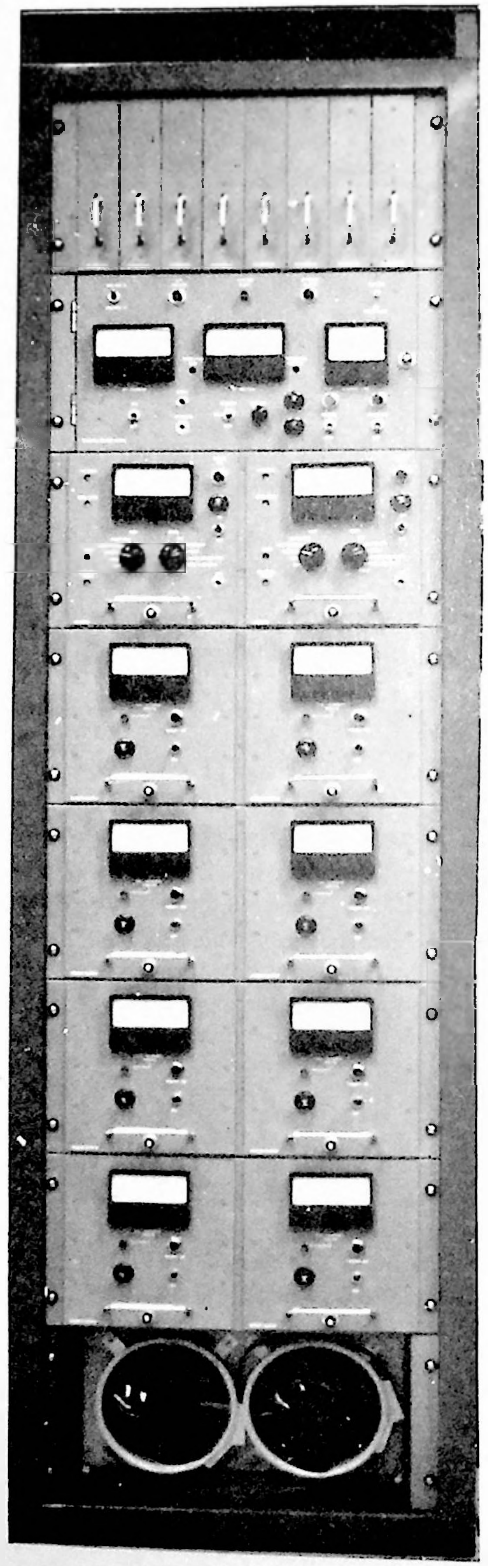
Imperative needs often spawn invention and the development of Radiobeacons has helped to confirm this by having spearheaded the inevitable thrust towards wholly solid-state transmitters of medium/high power capability with their theoretical potential of greatly improved reliability. From this early thrust there has emerged some new technology, new concepts of transmitter design, new concepts of modular serviceability, and a clear understanding of the pitfalls to be avoided in solid-state designs. It is now inevitable that much of this will spill over into the design of AM Broadcast transmitters and indeed, has begun to do so. This paper, accordingly, gives a particular historical perspective to this trend which may provide some insights into the future generations of Broadcast transmitters.

1969/70 - A SOLID-STATE LANDMARK

Early in 1969, the Canadian Transport Department came out to tender for some Aerobeacon transmitters at the 500 Watt carrier level. To test the state of the art, they stated in the technical documents that preference would be given to a wholly solid-state design. Several companies responded, but only one (NAUTEL) offered wholly solid-state. The others indicated that this was beyond the state of the art at that time.

Two choices were offered in the Tender response. One was for a fully duplicated system to give conventional standby back-up, but the other was for a single system with a "soft-failure" characteristic. Transport, despite having specified the former, were persuaded by the technical arguments in favour of the latter and chose it. By 1970, the world's very first medium powered, solid-state MF transmitter was a reality and soon thereafter was on the air in northern Newfoundland. A further 29 transmitters of this type were delivered over the following six months or so, and as they were installed in widely dispersed locations, a unique network of solid-state transmitters began to operate 24 hours per day without interruption. These transmitters have been running ever since and have provided startling, unshakeable evidence of the wide margin of superior reliability available from solid-state with the appropriate technology.

It is pertinent to mention at this point that, before this field evidence was in, there were many engineers (even within Transport) who issued dire predictions for the fate of "tubeless" transmitters. Their pessimism was proven unfounded and, from this early beginning 13 years ago, more than 4000 totally solid-state Radiobeacon-related transmitters have now been manufactured and installed



Vintage "70"
500 Watt Radiobeacon

throughout the world. For new Radiobeacon procurements, tube-type transmitters have now become the rare exception rather than the rule.

SOFT FAILURE

In the foregoing, reference has been made to "soft failure". This is a very important concept and will be described in more detail. A convenient analogy to facitate understanding of this principle is to consider a high power light source corresponding to, say, a 1 kW incandescent lamp. This light source might be provided by a single 1 kW lamp (if it were available) or by a cluster of ten 100 Watt lamps. The single lamp solution would produce catastrophic failure (total darkness) if the lamp failed whereas the ten-lamp solutiuon would merely produce reduced illumination if one lamp failed. In one case, repair is imperative whereas in the other it can be performed whenever convenient. It is important to understand that the "soft-failure" characteristic of the ten-lamp solution is only possible because the lamps all operate independently or in isolation. If now instead of ten lamps it was required to combine the outputs of ten transmitters, then this isolation would not occur without special circuit arrangements to ensure it. For example, if the ten transmitter outputs were simply paralleled, a short circuited output on any one would disable the remainder.

The remedy for this particular problem lies in the use of hybrid transformer networks originally developed many years ago for the telephone industry. Figure 1 shows an example of a modern hybrid network suitable for combining any number of RF sources with complete mutual isolation.

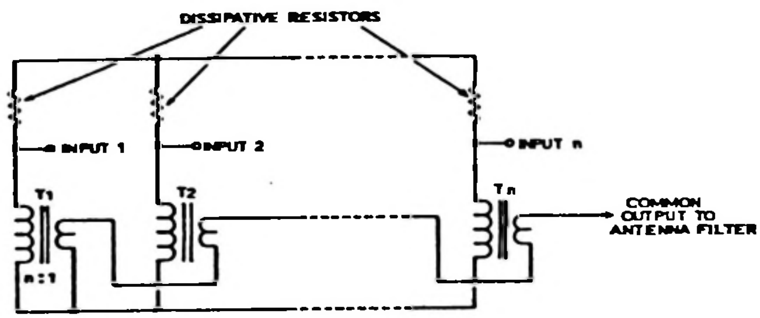


Figure 1 - Isolating Power Combiner (US Patent 4,156,212)

This type of network will combine the final outputs of several small transmitter amplifiers which is the most difficult problem. A further problem remains, however, in the need to coherently drive these amplifiers and supply them with DC power whilst maintaining failure isolation. Many arrangements are possible to achieve this, depending upon the required degree of isolation, and Figure 2 shows one such arrangement currently used by NAUTEL in their AMPFET transmitter series.

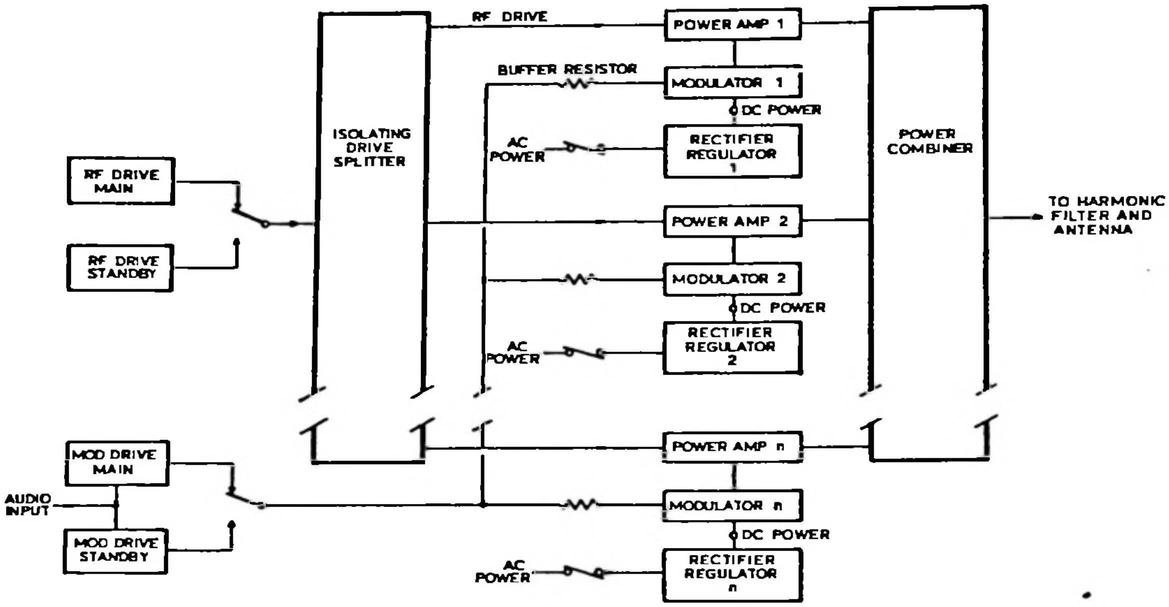


Figure 2 - AMPFET Block Details

It will be noted that the Modulator and Rectifier sub-systems are fully separate and thereby isolated. On the other hand, the Modulator Drive and RF Drive sub-systems have to be common and are therefore duplicated in the conventional Main/Standby fashion. If a fault in either of these latter sub-systems occurs, then it is sensed and automatic changeover to the Standby system takes place.

SWITCHING AMPLIFIERS AND MODULATORS

This topic has been extensively discussed elsewhere but will be included here for completeness. When the first high-power, solid-state Radiobeacons were introduced in 1969/1970, the technology for Class D power amplifiers and switching-regulator power supplies was already firmly established. To apply this existing technology to the new transmitters was an obvious step because of its inherent efficiency and basic digital simplicity, and accordingly, the second generation of solid-state Radiobeacons embraced these concepts.

Figure 3 illustrates the simple concept of the Class D principle.

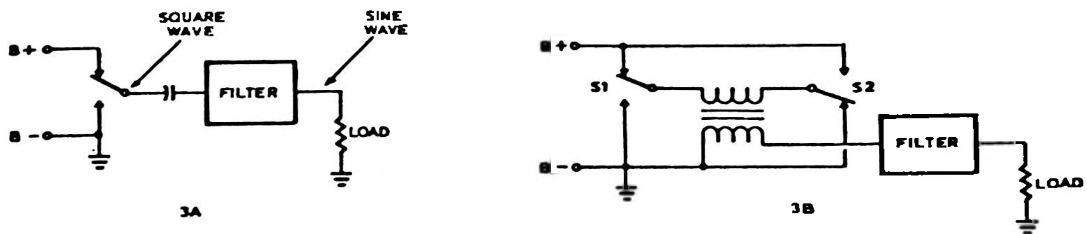


Figure 3 - Class D Principle

3a is a basic illustration whereby the switch S1 must be imagined to be switching back and forth at the carrier frequency. The resulting square wave output is then filtered to remove harmonic components. 3b is a push-pull version of 3a.

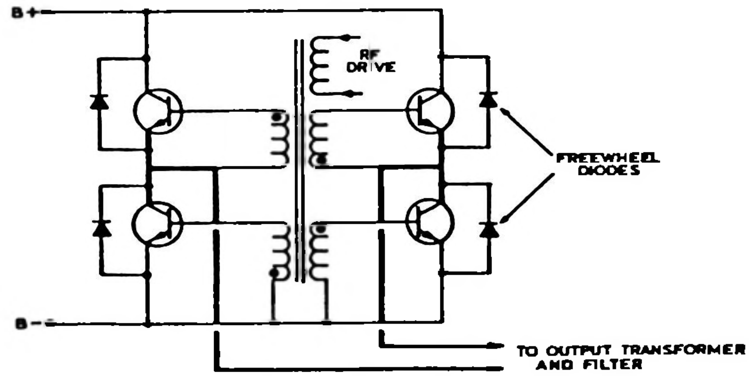


Figure 4 - Class D Principle Implemented with Bi-polar Transistors

Figure 4 shows the Class D principle implemented with bi-polar transistors. Note the freewheel diodes across the bi-polar transistors - these are most important if the circuit is to function with load SWR.

The standard switching regulator supply needs little further description here with its series switch, freewheel diode and filter (Figure 5).

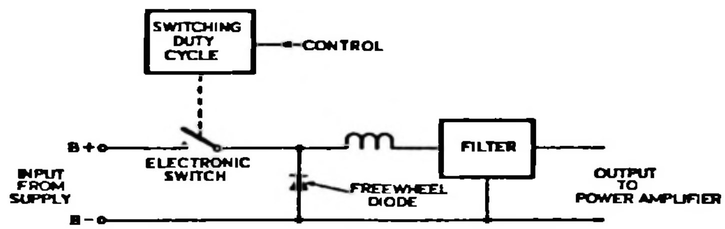


Figure 5 - Switching Regulator as Modulator

To use this as a switching modulator (pulse duration or duty cycle modulator), the control voltage for the switching duty cycle has both a DC and an AC component. The DC component determines the mean duty cycle and, hence, the transmitter carrier level, and the AC component (incoming audio) determines the modulation depth. Design of the filter requires more care than for a non-modulating supply because it must pass all audio frequencies yet reject the switching frequency and its side spectra. A multi-pole, low-pass filter with a notch at the switching frequency is commonly employed for this. With these switching techniques, a power amplifier efficiency of about 90% and a modulator efficiency of about 95% are readily available when using modern power FET devices as the switches, and the resulting system is fundamentally linear without any need for linear transfer characteristics or negative feedback.

SEMICONDUCTOR PROTECTION

The design of solid-state transmitters does require meticulous attention paid to the operation and protection of the transistor devices therein. A hermetically-sealed, silicon transistor is an incredibly reliable component providing that it is used correctly. However, a transmitter connected to an antenna is a fertile environment for misuse and failure to appreciate this fact has produced some "lemons" which have given a wrong impression to the market place.

Radiobeacons operate, for economic reasons, into very poor antenna which are much too short and require critical loading with a series inductance. The resulting system is sharply resonant, prone to environmental de-tuning and, with the very high antenna voltages present (50 kV typically), it is also very prone to antenna spark-over. Hence, designing Radiobeacons is a good place to start for worst-case experience. In contemplation of these conditions, NAUTEL originally adopted and have continued with the following standard design principles:

- A transmitter shall not damage or unduly stress its components when operating at any value of antenna SWR including open or short circuit.

- This SWR must be considered capable of instantaneous change.

- Adequate protection must be provided against lightning strike.

- A transmitter must be protected from the effects of sustained moderate power line overvoltages and short-time, large power line transients.

To this list should be added the "motherhood" but often overlooked design considerations, viz:

- With the worst-case antenna loading, power line voltage and so on, transistor junction temperatures must remain within limits which assure reliability (not maximum permissible temperatures).

- Only semiconductor devices and other components with assured reliability levels should be used (this normally requires hermetically sealed semiconductors and MIL type components).

- Internal sensors should continuously monitor for device stress either from antenna loading or modulation over-drive and take instant protective action.

The impressive field data from the 13 years of operation of the early NAUTEL Radiobeacons can be interpreted to confirm the validity of these principles. In contrast, some attempts to transplant standard tube transmitter technology to this field have produced dismal results.

ANTENNA SWR

This topic deserves some separate discussion. In the field, the transmitter load impedance hopefully is reasonably correct with low SWR, but occasionally, due to ice, snow and even temperature change, the antenna SWR might become temporarily unacceptable. Unacceptable, that is, for the Class D amplifiers to continue to operate at full load current without stress. One remedy would be to shut the transmitter down when this occurs, but this remedy would create a nuisance factor and a far better remedy is to reduce the transmitter power until the SWR problem has passed. This solution can be accomplished by what is referred to as "reflected power AGC". With this, a voltage taken from the reflected power terminal of the transmitter reflectometer is used in a feedback AGC fashion to control the transmitter power level so as to always maintain non-stressed operation. Special circuits have been developed to perform this control function without introducing any distortion.

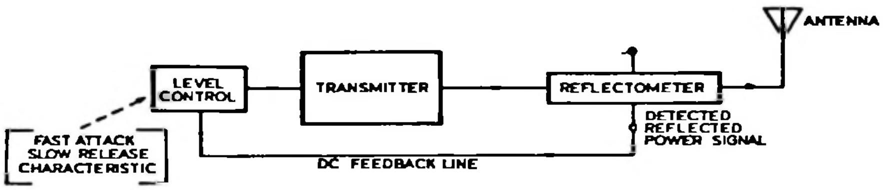


Figure 6 - Reflected Power AGC

MODULES AND SERVICING

It was stated earlier that Radiobeacons are often installed in remote, difficult-access locations. These sites are unattended and might only be visited on an annual basis, and for this reason any fault identification and remedial action for a defective transmitter should be as easy as possible. Fortunately, the solid-state transmitter with its multiplicity of similar sub-systems (modules) can be designed to streamline this procedure. If it is assumed that internal sensors and "fault lamps" pinpoint a defective module and further that the modules simply plug-in via the front panel, then servicing can be as simple as changing a light bulb. Most Radiobeacons now have these features and the serviceman carries some spare modules with him and can then repair any defective ones at his leisure back at the depot. For the serviceman who has several transmitters of different power levels in his charge, a further bonus occurs in that they will all use the same basic modules, so simplifying his spares inventory.

OTHER FEATURES

A number of miscellaneous features sharply differentiate MF solid-state transmitters from their "tube-type" forebears. These are conveniently expressed in the following summary:

- Solid state transmitters generally require just a single, low voltage power main supply of about 70 volts or so. This is inherently safer than the high voltage supplies required for tube transmitters with a corresponding reduced risk to service personnel. The AC power transmformer is obviously simpler and can readily be designed for virtually zero failure probability (of the approximate 2000 Radiobeacons delivered from the U.S. and Canadian plants of NAUTEL, not a single main power transformer failure has occurred in the 13 years).

- The filament warmup - bias voltage - plate voltage, interlocked sequence of tube transmitters is eliminated. Switch-on can be instantaneous if so required and remote control is accordingly simplified.

- The solid-state transmitter is generally much smaller than its tube counterpart, generates less waste heat, and consumes much less primary power. All these factors, especially the last one, can result in substantial user economies.

- Solid-state devices do not exhibit the aging and wear-out characteristics of tubes. If, for example, a solid-state transmitter will deliver 50 kW peak power upon installation, then it will continue to do so indefinitely.

- Solid-state designs are inherently wide band, highly linear and will accept AM stereo drive signals without any problems.

RECENT DEVELOPMENTS

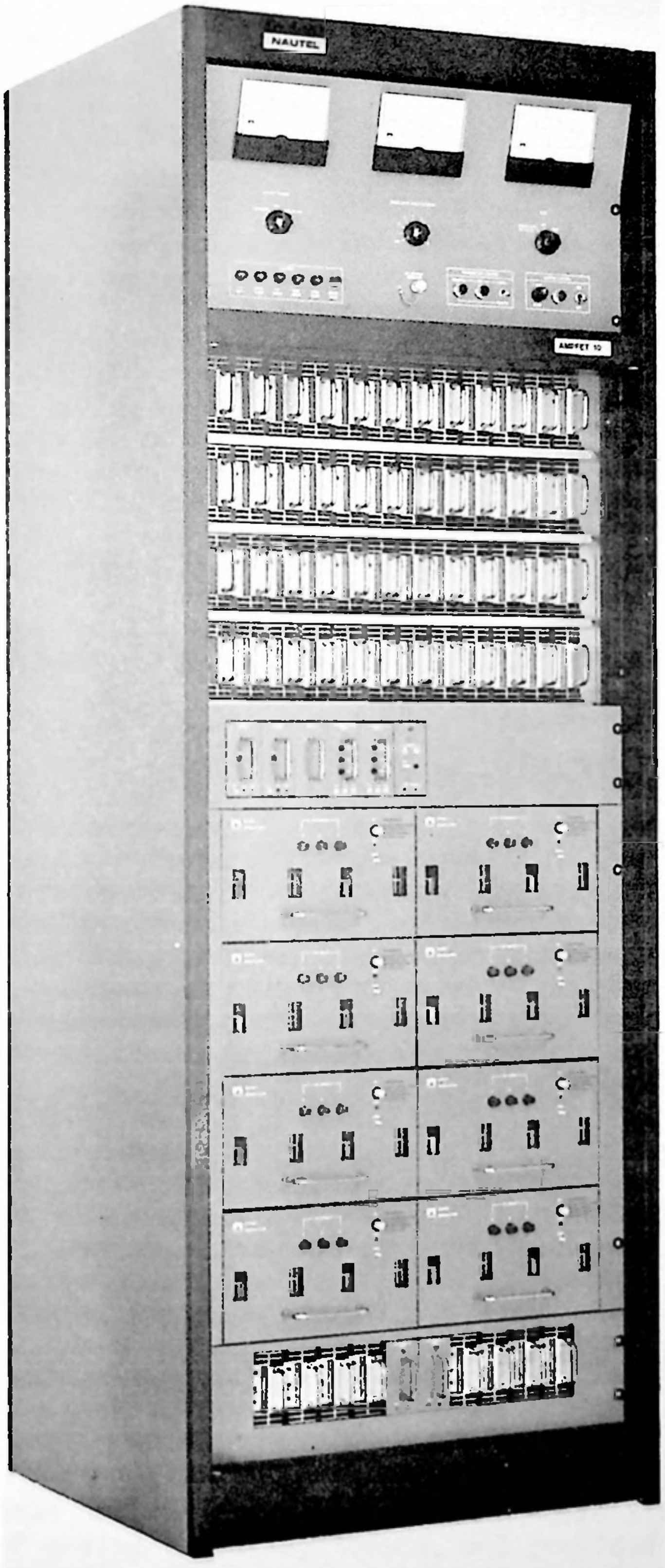
The features and characteristics of the Radiobeacons so far described were hardened more than a decade ago. Over this decade, small innovations and techniques have served gradually to improve the product and streamline the production process, but a more significant step forward occurred 3 - 4 years ago. This step was the use of power FET's in place of the bi-polar transistors used up to that point. The solid-state transmitter designer had long awaited a better switching device than the bi-polar transistor (with its switch-off limitations following saturated switch-on), and was delighted when suitable power FET's became available. Their early promise was quickly verified and a substantial increment of efficiency, reliability and economy became possible. Once again, there were pessimists who cast doubts on these new devices, but they were quite wrong and the author confidently predicts that power FET's will rapidly displace bi-polar devices for PA and modulator switching elements in new MF designs.

TECHNOLOGY TRANSFER

With the significant efficiency and economic advantages conferred by the power FET and with the growing concern relating to the cost and the desire to conserve energy, NAUTEL decided in 1979 to extend their experience in the design of Radiobeacons to the AM Broadcast market place. By late 1980, the AMPFET series of Broadcast transmitters with 1 kW, 5 kW and 10 kW models were prototype realities and some broadcast engineers were invited to preview these in the laboratory and offer their comments. These comments were valuable and taken into account in the final production designs and specifications. Production of the new models commenced in late 1981 with first deliveries commencing in 1982.

The world's very first operational 10 kW totally solid-state Broadcast transmitter was delivered to CJFX of Antigonish, N.S. and commenced operation on March 15, 1982. Some minor teething problems did occur, as might be expected, but were quickly remedied with negligible down time. At the time of writing this paper, the Antigonish transmitter has almost completed its first year of operation and CJFX General Manager, Mr. David MacLean offers the following verdict: "The year's operation has been most successful. I am pleasantly pleased with the surprisingly small incidence of teething problems, the reduced power bills, the virtual absence of maintenance costs, and the freedom from difficulty in procuring replacement tubes. The transmitter has delivered exactly what was promised."

Other AMPFET transmitters at various power levels have also been delivered and installed more recently with a gathering momentum and Radiobeacon technology can be said to have now truly entered the AM Broadcast field.



Vintage "82"
AMPFET 10 kW Broadcast Transmitter

THE FUTURE

No article of this kind today is complete without a little prophecy regarding the future. This one is no exception and the author wishes to conclude it by stepping into this risky arena with the following opinion statement: "Solid-state has now virtually displaced vacuum tubes for useages such as radio receivers, television receivers, computers, and so on and has conferred very significant advances to these fields by so doing. This displacement process has now also occurred in the Radiobeacon transmitter field and will continue into the AM Broadcast transmitter field at <u>all</u> power levels. Credibility and a natural inertia to resist change will slow this process, but by 1985 at least 50% of new AM transmitter procurements will be solid-state and by 1990, this figure will be at least 90%."

ELECTRICAL AND MECHANICAL ANALYSIS OF

SYNTHETIC TOWER GUYS

Leopold Gregorac

Radiotelevizija

Ljubljana, Yugoslavia

and

Gregory F. Bowen

Philadelphia Resins Corporation

Montgomeryville, Pennsylvania

INTRODUCTION

From a theoretical point of view, the guying of radiating towers has always been considered a very complex problem. When higher radiated power levels are involved, serious practical problems can arise. Careful attention must be given to electrical and mechanical aspects when designing any guyed transmitting antenna in LF, MF or HF fields.

In the past the problem has typically been solved by sectioning standard steel guys with the insertion of ceramic insulators. In some cases very complicated metal armatures have to be applied. At times the complexity of properly insulating guy wires leads some producers to prefer self-supporting towers even though larger tower cross-sections result in inferior radiating characteristics.

About ten years ago, synthetic tower guys were put on the market which had good insulating properties. At that time, they represented a promising method of guying radiating towers, and provided possible solutions to numerous problems including flashovers, maintenance, reduced interference, etc.

Unfortunately, the utilization of synthetic tower guys was not carefully analyzed ten years ago. Broadcasters who used synthetic guys on radiating towers of higher power for the first time were faced with difficulties and were often disappointed in performance. Producers, however, were not discouraged and continued to improve their products.

In 1976, a 10 kW medium wave umbrella antenna guyed with synthetic tower guys was put in trial operation by RTV-Ljubljana. We felt proper electrical protection of the synthetic guys should be considered carefully and we set about studying the problem.

Results of our two-year study and experiments have been considered in guying a set of towers that are presently operating. The results have been published in article (1). The purpose of this paper is to discuss additional developments resulting from studies of the behavior of synthetic tower guys in the broadcast environment.

SOME TYPICAL PROPERTIES OF A GOOD SYNTHETIC TOWER GUY

A synthetic fiber rope designed for use as a tower guy consists of a fiber core carrying mechanical forces and an extruded jacket which helps protect the core against electrical, atmospheric and other conditions affecting it. Let's examine some properties of the synthetic fiber rope, PHILLYSTRAN® HPTG, produced by Philadelphia Resins Corporation.

Core Material	Kevlar® Fiber
Jacket Material	Olefin Copolymer
Diameter, Jacketed Guy	15.5 mm (0.61 in)
Approximate Weight, Jacketed Guy	174.1 kg/km (117 lbs/1,000 ft)
Minimum Break Strength	120 kN (27,000 lbs)
Tensile Modulus of Elasticity	18 x 10^6 psi
Operating Temperature Range	-55°C to 80°C (-131°F to 176°F)
Resistance Against Corrosion	100% Noncorrosive
Flammability	Self-Extinguishing
Water Absorption, Core	<1%
Water Absorption, Jacket	<0.12%
Dielectric Constant	2.6
Dielectric Strength	19.7 kV/mm
Dissipation Factor	$D = 1.9 \times 10^{-3}$

Comparative stress versus strain curves are given for PHILLYSTRAN® HPTG and other guy material candidates in Figure 1. Also, jacketing materials can be applied with a spiral or helical shape (Patents pending) in order to reduce the probability of vibration and enhance self-cleaning in high pollution areas. This is designated PHILLYSTRAN® HPTG-NA (non-aeolian) by the manufacturer.

PHILLYSTRAN® guys are connected at each end with specially designed end fittings (Figure 2). After removal of an appropriate length of jacket, the Kevlar® fiber core of the guy is opened and spread. The end fitting is slipped up over the opened section of PHILLYSTRAN® and filled with resin socketing compound to complete the assembly.

MECHANICAL ASPECTS OF SYNTHETIC FIBER TOWER GUYS

General

If a tower is already guyed with steel guys, these may be replaced directly with synthetic guys without essential modification to the tower itself.

Considering the risk of brush fires, vandalism, etc., a 20 ft. long steel lead line is recommended at ground level followed by the synthetic fiber guys onward up to the connecting points on the tower. Standard guy tensioning equipment is then easily utilized as well.

End Fittings

Extensive testing has been done in order to optimize the shape of resin socketed end fittings used with synthetic guys (4). In order to achieve proper break strength at the end fitting/guy interface, it is important to choose a relatively long tapered internal socket shape with a length-to-guy diameter ratio of at least 6:1. The angle of the taper with respect to the longitudinal axis of the end fitting should lie between 3-4° in order to minimize internal stress (Figure 2a).

In order to maintain the lightweight nature of synthetic guy assemblies, aluminum alloy with high tensile properties was chosen as the end fitting material. End fittings have been successfully fabricated from aluminum bar stock and by casting methods.

The electrical characteristics of the designed end fittings are discussed later in this paper (Patents pending).

Guys

In order to avoid excessive tower deflections, it is important to use high tensile modulus materials such as Kevlar® fiber for the load bearing portion of the guy. In addition, synthetic fibers must be chosen which exhibit the absence of long-term static creep. Kevlar® has been extensively tested for this characteristic and provides for a final guy which will not require retensioning.

The rope construction utilized in the manufacture of synthetic tower guys must be chosen to minimize constructional stretch and maximize the inherently high modulus found in the Kevlar® fiber. A parallel or close to parallel construction allows the finished guy to reflect the high modulus of elasticity available from the Kevlar® fiber components.

The vibration damping capacity of synthetic fiber ropes is rather low and may allow excessive vibration of the complete structure when the wind blows. There are several vibration attenuating methods available. In some instances simply maintaining guy tensions at 5% of minimal break strength has reduced vibration. Tower deflections will not become excessive at these low guy tension levels because synthetic guys are five times lighter than steel guys. A nearly undetectable catenary would be created in a synthetic guy which is reduced from 10% of minimum break strength to 5% when compared to steel guys. However, tower manufacturers should be contacted regarding reduced tensions with synthetic guys.

A spiral shaped extruded jacket can be used in conjunction with synthetic guys. PHILLYSTRAN® HPTG-NA [non-aeolian (Patents pending)] is an example (Figure 2b). The spiral shape helps to break up vortices developed behind a given guy when the wind blows and reduces the tendency toward vibration.

Finally, there are a wide assortment of vibration dampers on the market which have traditionally been used to dampen aeolian vibration in horizontal electrical conductors or in steel structures. These devices are equally applicable to synthetic fiber guys.

ELECTRICAL ASPECTS OF SYNTHETIC FIBER TOWER GUYS

As suggested above, the mechanical properties of well designed synthetic fiber ropes are equal to or better than those of steel ropes. The overriding reason, however, that radio tower guys made of dielectric material have been put on the market is their effective insulating property in the electric field. Let's examine their response when analyzed in an electric field.

Radiating Towers

A radiating tower is mainly used as an antenna in LF, MF and HF fields. Transmitters have already reached power levels as high as 2000 kW in LF and MF fields. As we get closer to the tower, electric field strengths become more intense. Table 1 below gives the maximum electric field strengths occurring on normal towers which are mounted on insulators. Different heights are considered with a power level of 2000 kW. Transmission line equations, which are exact enough for this purpose, have been applied to evaluate them.

Table 1
Maximal Electric Field Strength for 3 Typical Towers (peak values)

P = 2000 kW m = 100%

height	h...height		λ...wavelength
	$h = 0.1\lambda$	$h = 0.25\lambda$	$h = 0.5\lambda$
radius	$\|E_{max}\|$ [kV/m]	$\|E_{max}\|$ [kV/m]	$\|E_{max}\|$ [kV/m]
200	1610	56	54
100	805	33	29
50	376	20	10

As illustrated in Table 1, the maximum electric field strengths occurring are less than 2 kV/mm. Considering that PHILLYSTRAN® synthetic tower guys have a dielectric strength of 19.7 kV/mm, there is a sufficient safety factor. This is valid only when the weather is fine and undisturbed. Moisture, polluted and salted air, etc., each in their own way decrease the insulating capacity of the synthetic guys. Theoretically, it is impossible to find out to what extent dielectric strength is actually reduced. As far as moisture is concerned, the above described synthetic guys have a jacket which has very low water absorption. In addition, socketed ends are sealed with silastic adhesive in order to minimize the possibility of internal accumulation of water. The completed guy materials are considered to be a good quality insulator (2). Accumulated field experience and experiments performed during the three-year studies confirm the excellent behavior of synthetic guys in wet atmospheres.

It is more difficult to analyze synthetic tower guy resistance to leakage currents. Polluted air creates dirty spots on the jacket which may become conductive. Very intense electric field strengths may cause flashovers between two conductive spots on the surface of the guy. It is possible that enhanced conductivity between spots may gradually burn the rope.

We have achieved a higher resistance against leakage currents by providing a spiral shaped jacket surface, which lengthens the path for leakage currents. In addition, a silicone boot has been added to the end fitting assembly at the most critical point where the synthetic guy enters the end fitting (Figure 2a and 2b).

VHF and UHF Broadcasting Towers and Synthetic Guys

From an electrical point of view, the application of synthetic guys for VHF and UHF installations has proven very useful. PHILLYSTRAN® HPTG guys are practically transparent in the electric field and do not disturb the radiating diagrams of VHF and UHF antennas even if synthetic guys are situated in front of them. In these applications, maximum powers of transmitters are in the range of 40 kW. In practice, this power is usually distributed among several transmitting antennas so that input power of one is a maximum of 3 kW. Considering the gain of an antenna (approximately 10 times), its maximum radiated power is in the range of 30 kW. Electric field strengths occurring in the near field of one antenna or the far field of multiple VHF or UHF antenna arrays are much lower than those which occur near the tower on MF antennas. Synthetic tower guys are completely appropriate for use on VHF and UHF antenna arrays.

Behavior of Synthetic Tower Guys in the Earth's Electrostatic Field and Their Response in a Thunderstorm

As of the date of this paper, no direct strike of lightning on synthetic tower guys has been witnessed by us. This is due to the inherent shape of towers which protect the ropes against lightning. It is known (3) that a protected region exists in an area transcribed by 45° angle from the top of a given tower.

Another important consideration is the influence of the Earth's electrostatic field. In the near vicinity of a thundercloud and in certain weather conditions, particularly if a site high above sea-level is in question (2), the electric field strength may achieve the value of 20 kV/m. The electrostatic field fluxes concentrate around the tower, grounded for static phenomena, as illustrated in Figure 3. The maximum electric field strength is always on the tower while the direction of the field's vector is normal to the surface of the tower. Illustrated in Figure 4 are the values of electric field strengths for 200 meter tall towers with different cross-sectional dimensions. Electric field strength may be even more intensive on sharp and rough metal edges on or near the tower (1) and corona effect may be a result. If edges of potting heads are sharp and rough, the corona effect may occur very close to the synthetic fiber rope. In fact, the energy of electrostatic field is usually not very high (2). This prevents the rope from immediately burning through, but separate little flashovers may, to a smaller extent, affect the insulation. Moisture and polluted air aggravate the situation. Provided a radiating tower is involved, the energy radiated by the tower burns through the rope at the point where corona effect from the electrostatic field occurs. So far, this has happened in practice, when proper electrical protection of synthetic tower guys has not been used.

Taking into consideration all the above facts, it is logical to conclude that the most exposed areas need to be appropriately protected. This protection may be performed in several ways. First and above all, the most exposed point where synthetic fiber rope enters the potting head must be protected. By using a smooth enlarged funnel-like end fitting, as illustrated in Figure 2, electric field strength is decreased by 8 times in comparison to standard sharp conical end fittings (1). The possibility of corona effects is thereby minimized.

If higher transmitting powers or complex antenna systems are used--where electric field strengths of extreme intensity occur--additional protection has to be provided, especially on the upper guys. A pair of ceramic egg insulators positioned between the corona end fitting and the tower leg is an effective protection (see Figure 5). In this way, electric field strength at the most exposed point of the synthetic guy is decreased by an additional factor of 4 times. Additional corona rings (see Figure 6) may be used for protection. By method of applying the Fredholm's integral equation of first kind, we have evaluated that by this method of protection electric field strength is decreased by about 3 times at the most exposed point. Figure 7 illustrates the differences in electrical field strengths around the various potting heads or end fittings. Appropriate levels of electrical protection should be specified when designing tower systems with synthetic guys.

CONCLUSION

Synthetic tower guys which are properly designed from a mechanical viewpoint and which are provided with proper levels of electrical protection are available today as complete assemblies and can be used safely and effectively for guying broadcasting towers. Our opinion is based on theoretical results and measurements from field experiments on towers ranging in power from 1-300 kW that are presently operating.

For effective application, the following criteria must be met:

--The core material chosen for synthetic tower guys must have a high tensile modulus, its dielectric strength should be the highest available and its water absorption coefficient should be the lowest possible.

--The jacketing material chosen for synthetic tower guys should be smooth but thick enough to protect the core against mechanical and environmental damages. Its dielectric strength should be the highest available and water absorption coefficient the lowest available. A spiral-shaped smooth jacket is recommended where high pollution levels are known to exist and high transmitting powers are involved so that the possibility of leakage currents and surface deterioration is minimized.

--In order to minimize aeolian vibration, which is more easily generated in lightweight synthetic tower guys, three alternatives are available. Synthetic guys can be tensioned at lower levels than steel guys because resulting catenaries are so small tower deflection is not affected. However, tower manufacturers should be consulted on an individualized basis. The use of spiral jacketing tends to break up the forces which cause aeolian vibration. Commercially available dampers can be attached to synthetic tower guys to dampen vibration.

--A properly shaped synthetic boot should be located around the synthetic tower guy in the critical electrical region where it exists from the end fitting. A properly designed and assembled boot will further protect the synthetic guy from the possibility of leakage currents. End fittings should be completely sealed with a suitable elastic adhesive so that water absorption by the overall assembly will be minimized.

--It is advisable to attach a 20 ft. steel lead line at the anchoring end of each guy so that risks of brush fires and vandalism are minimized.

--Synthetic tower guys must be protected electrically with the attachment of properly designed funnel-like corona end fittings.

--It is advisable to further protect synthetic tower guys with ceramic egg insulators or with additional corona rings. This is particularly necessary when high transmitting powers or complex antenna systems with extreme electric field strengths are involved.

REFERENCES

1. Gregorac, L., 1981, "Eine Analyse der elektrischen Gegebenheiten bei der Verwendung von Kunststoffseilen zum Abspannen von Masten", RTM Heft 5, 201-212.

2. Hahn, L., Munke, I., 1973, "Werkstoffkunde die Elektrotechnik und Elektronik", VEB Verlag Technik Berlin, 219-367.

3. Golde, R.H., 1973, "Lightning Protection", Edward Arnold Ltd., London.

4. Test performed by Technische Hogeschool Delft, Mr. L. Wiek, September 25, 1981, for EMHA Technisch Bureau b.v. PHILLYSTRAN® European Office, The Netherlands.

Printed with permission of Philadelphia Resins Corporation, USA and EMHA Technisch Bureau b.v. PHILLYSTRAN® European Office, The Netherlands.

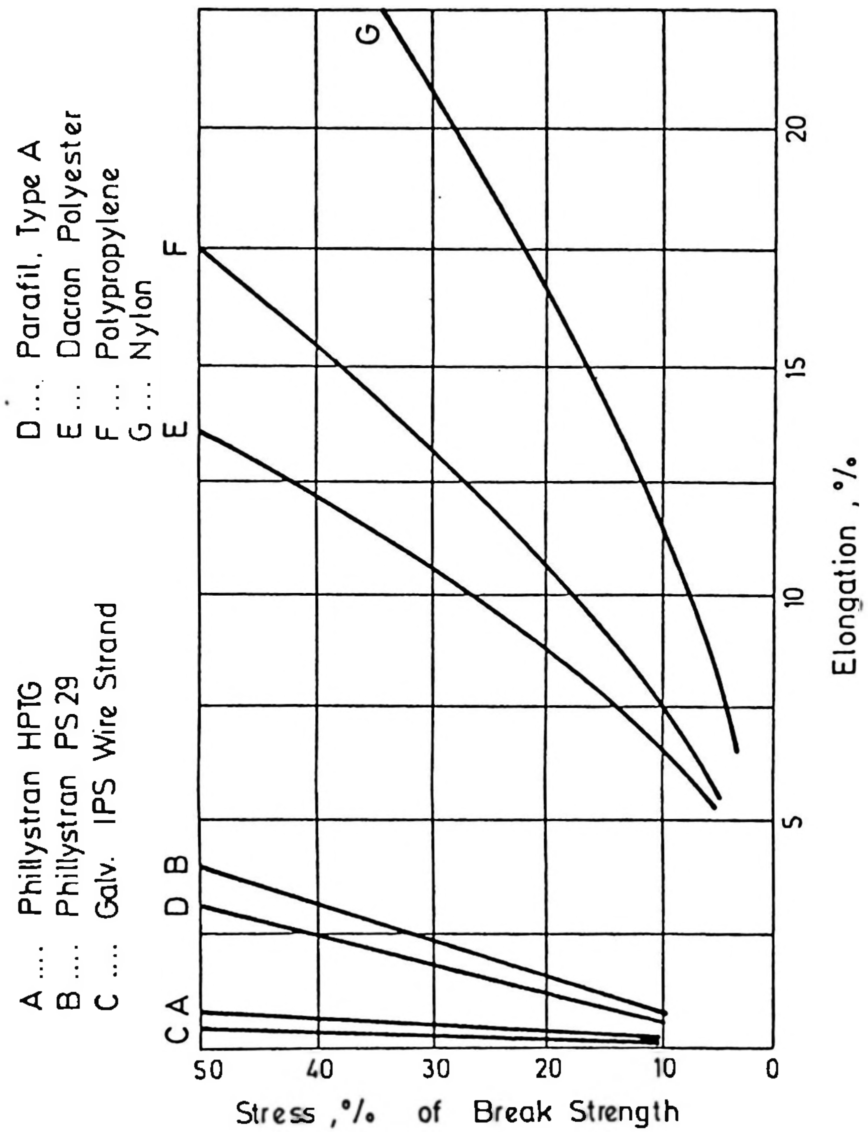


Figure 1 Comparative stress-strain characteristics of various ropes

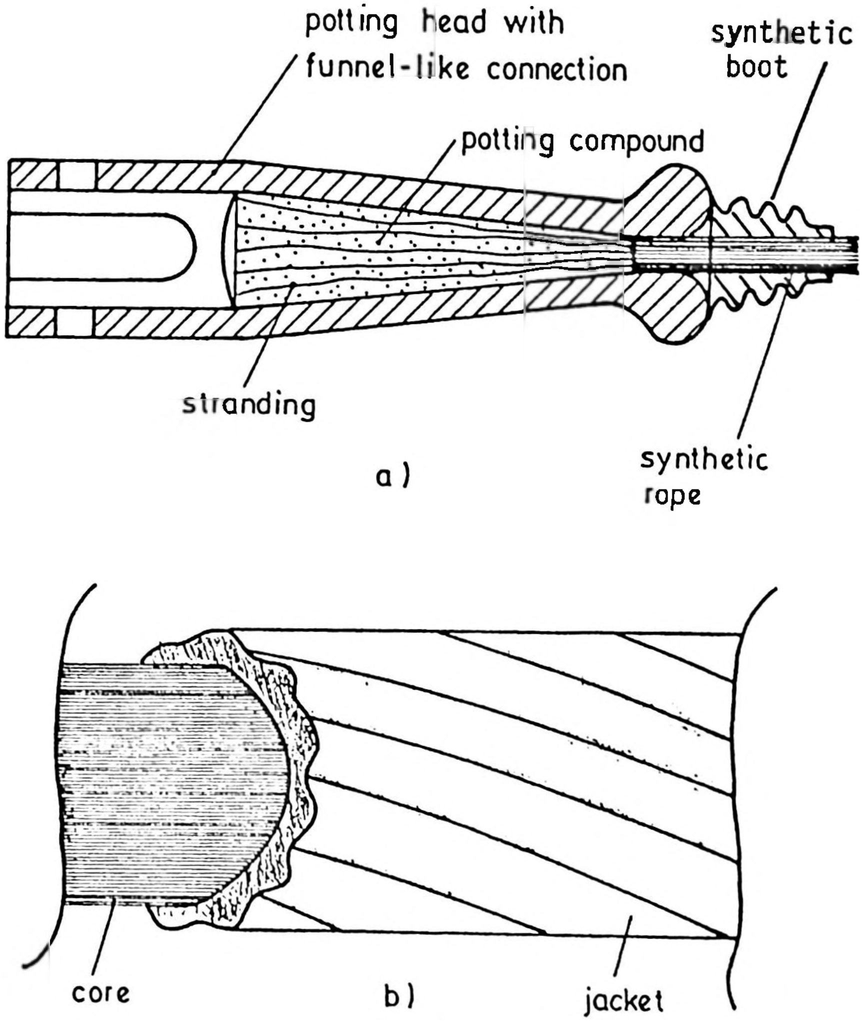


Figure 2 Potting head for synthetic rope (a) and synthetic rope with spirally-shaped jacket (b)

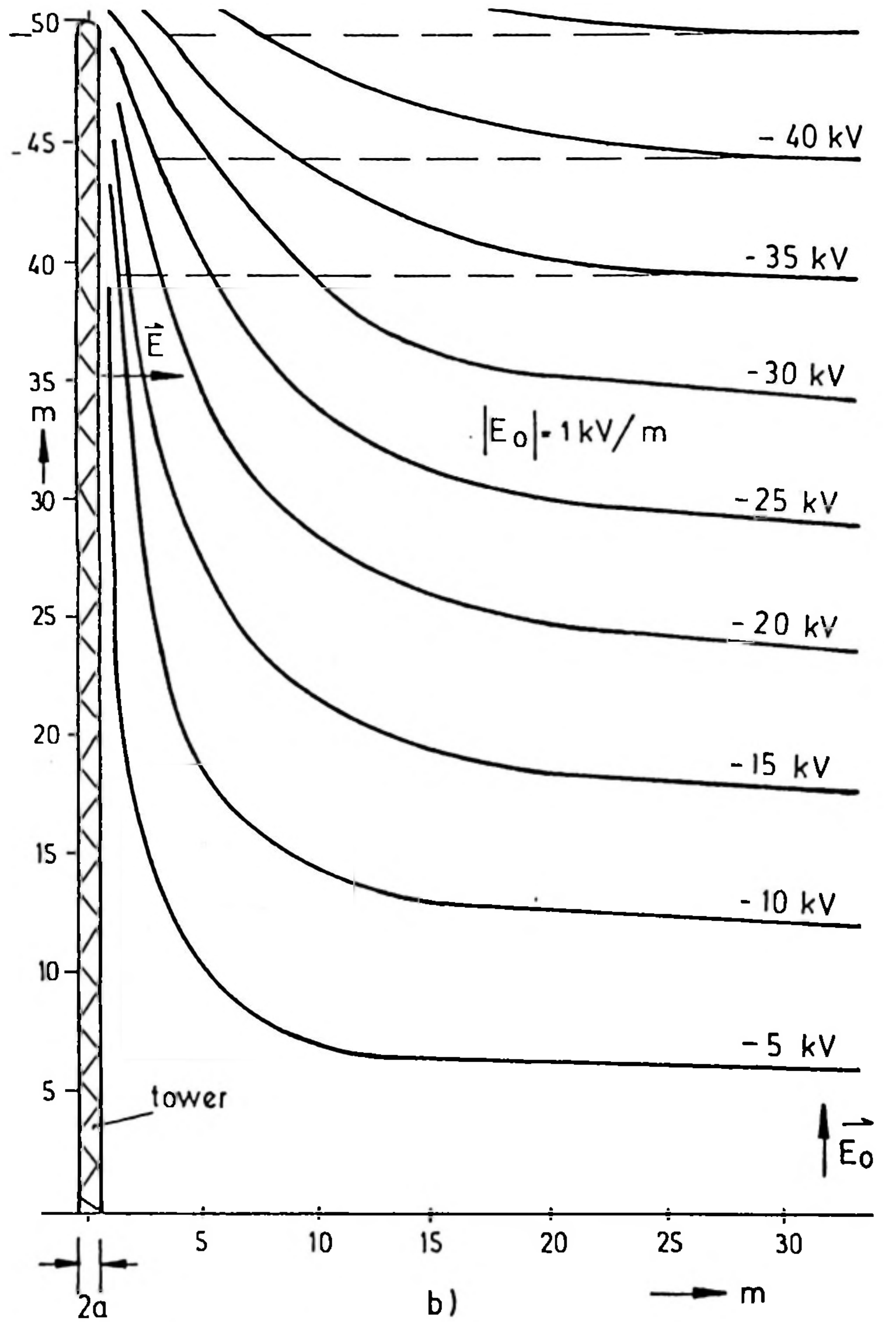


Figure 3 Equipotential lines around the tower caused by electrostatic field $\hat{E}_o$

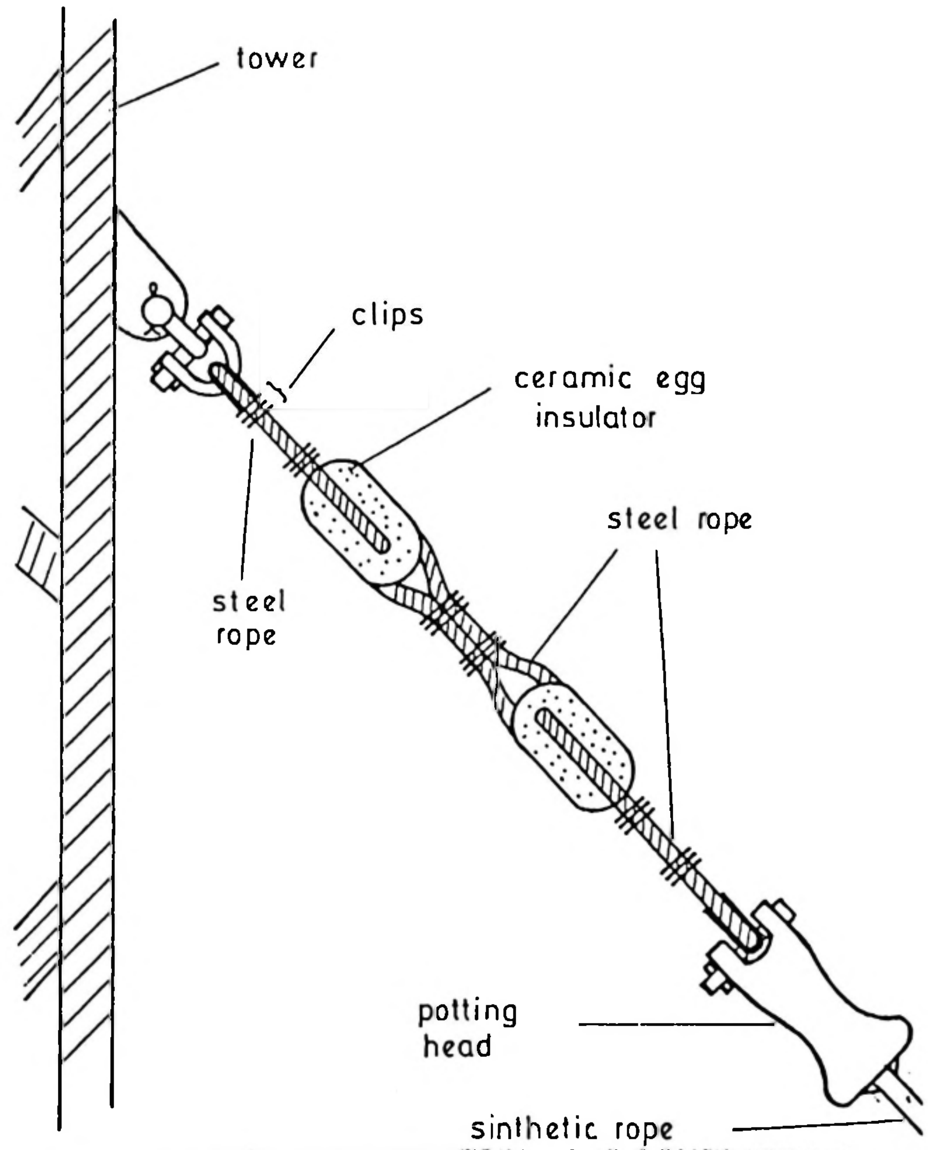


Figure 5 Additional protection by two ceramic egg insulators

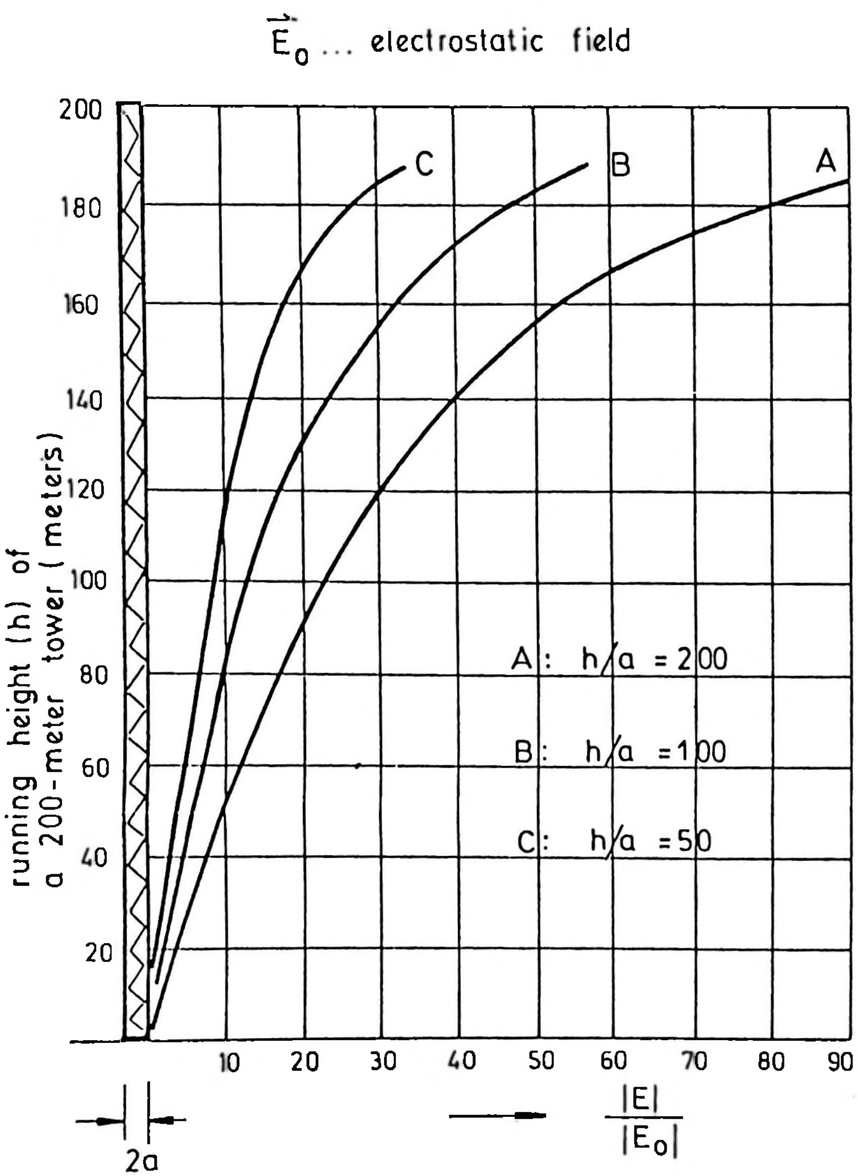


Figure 4 Electric field strenght E on the surface of a 200-meter tower caused by electrostatic field $\vec{E}_o$

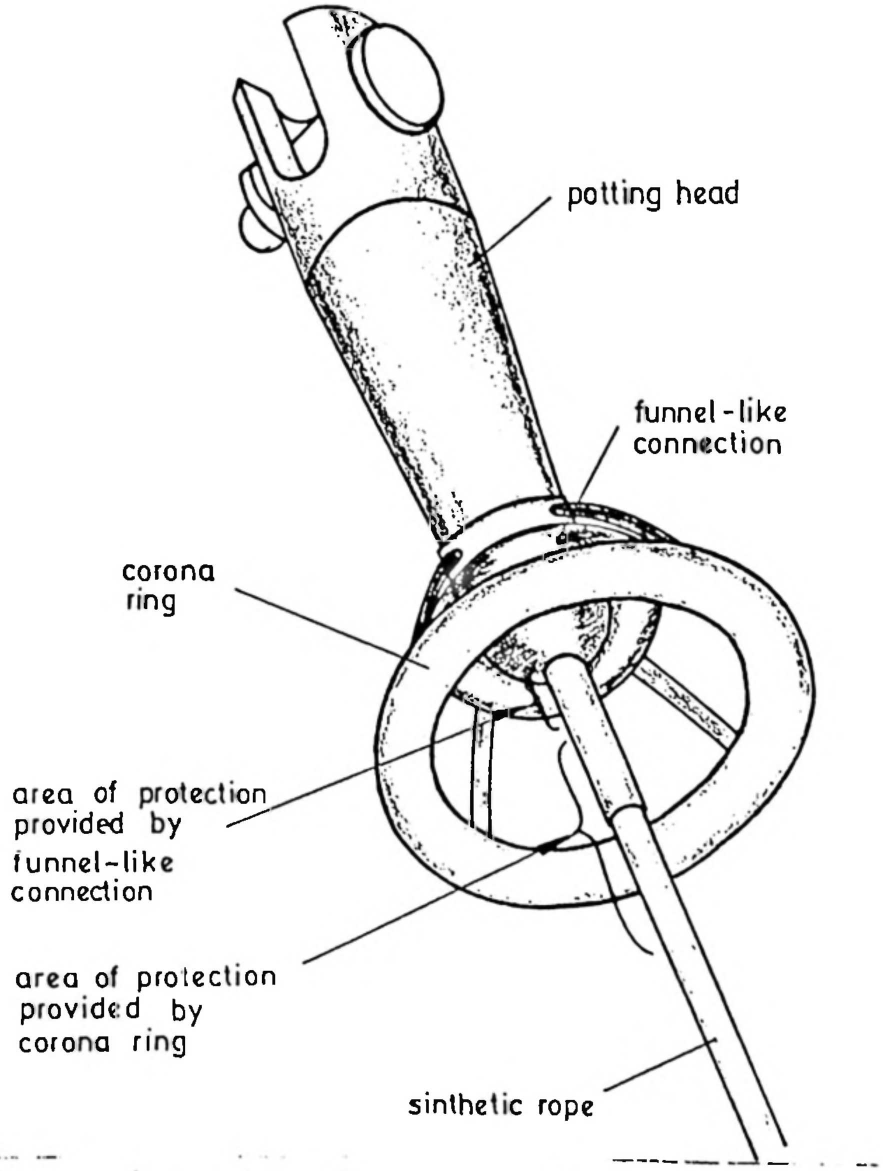
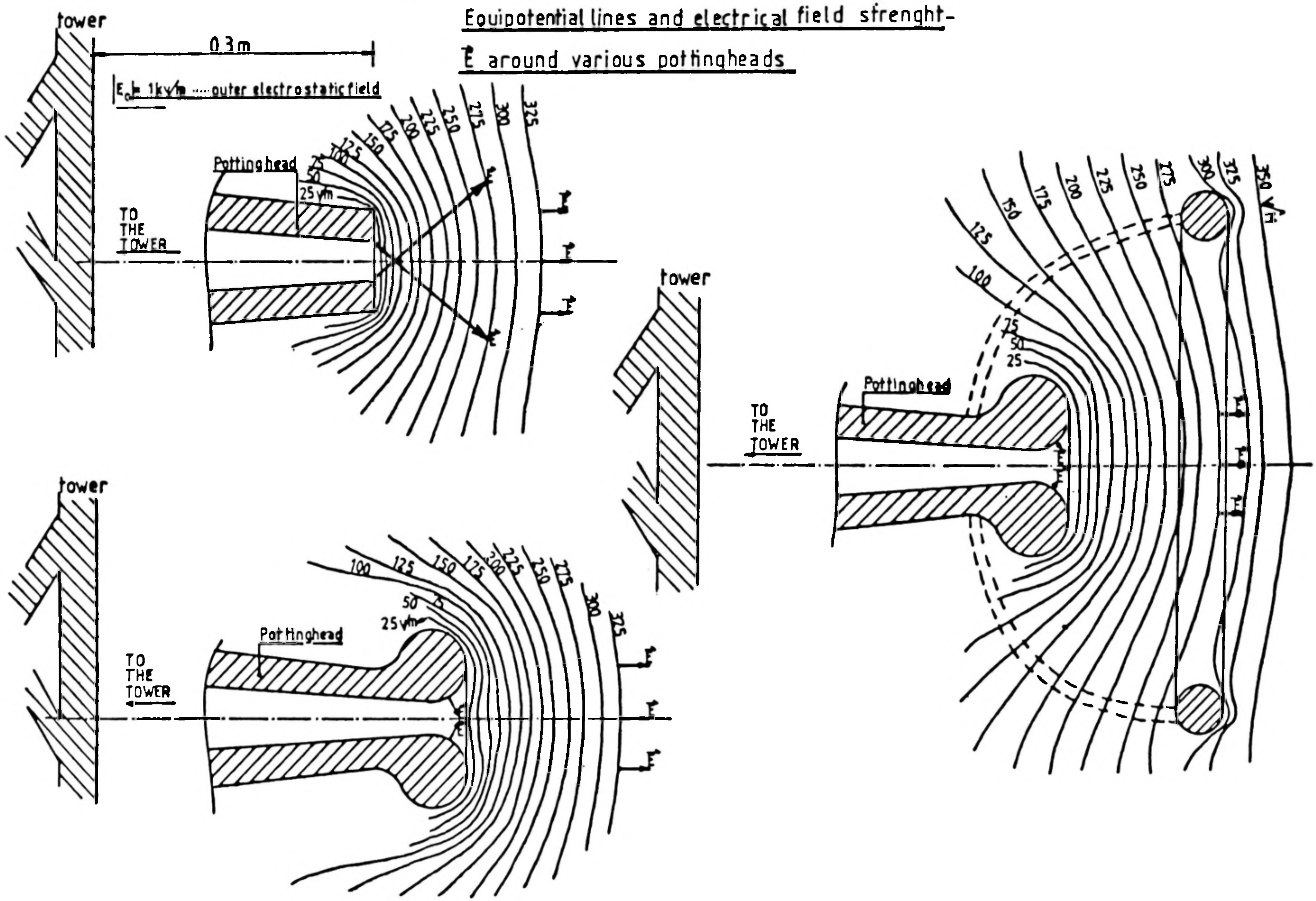


Figure 6 Additional electrical protection with an extra corona ring.

Figure 7:

Equipotential lines and electrical field strenght- $\vec{E}$ around various pottingheads



IMPACT OF REDUCED MILEAGE SEPARATIONS ON FM BROADCASTING

A Technical Report on Docket 80-90

Michael C. Rau

National Association of Broadcasters
Washington, D. C.

In a Notice of Proposed Rulemaking (Docket 80-90) issued on March 14, 1980, the Federal Communications Commission (FCC) proposed to adopt new classes of FM radio stations, modify minimum separation requirements, and permit "intermixture" of lower power assignments (Class A -- 3 kw) with moderate to higher power assignments (Class B/C -- 50 kw/100 kw) on the same channels. Although these changes are largely technical in nature and broad in scope, a great deal of useful technical information was not filed with the FCC in 1980 and therefore not included in its legal record. At the time, many industry engineers were preoccupied by the then red-hot technical issue of 9 kHz AM channel spacing. Recently, however, a resurgence of technical interest in Docket 80-90 has emerged and, as a result, five technical studies were performed during the past year and submitted to the FCC. The studies were filed by the "Subgroup on Technical Matters of the Advisory Committee on Radio Broadcasting", a Government-Industry collection of broadcasters, lawyers, and engineers who were interested enough in Docket 80-90 to supplement the Commission's record with useful and fundamental technical information. Each study, detailed below, addresses important and essential elements of this proceeding. The purpose of this article is to present the substantive results of these studies, and to show their relevance to Docket 80-90.

INTRODUCTION

Much of the research that was performed is original analysis of the concepts and standards which underlie the Commission's system of FM allocations. The work required substantial involvement, time and effort by many industry engineers who chose to supply the Commission's staff with technical support and information not otherwise available within the agency. There are two principal reasons for the industry's widespread interest. First, Docket 80-90 may be the last opportunity to address FM allocation technical standards while the FM spectrum can still support additional stations. If the FM spectrum becomes exhausted, any future alteration of allocation standards would be essentially meaningless. Second, the technical integrity of the FM broadcasting must be maintained. Degradation of the FM service now available disserves the public interest. Moreover, it is the Commission's responsibility, through its policies, standards and procedures, to determine the permissible degree of interference within the structure of FM allocations. It cannot be handled outside the agency: interference among broadcasters is a matter which cannot be left to the "marketplace" and is beyond the control of individual station licensees.

The Commission, therefore, needs to have a comprehensive, broad-based, technical record within Docket 80-90 if the implications of altering allocation technical standards are to be fully evaluated. It is this view which has prompted industry research. A genuine sense of scientific discovery pervaded the subgroup discussions as answers to important technical questions, some of which had been pending for over 30 years, began to materialize. In the words of one participating and helpful engineer, "we are seekers of the truth."

Summary of Technical Studies

The studies together with principal conclusions are listed below.

1. FM Broadcasting Receiver Characteristics and Protection Criteria (filed July 7, 1982). This report investigated four studies of FM receivers and determined the implications of receiver characteristics on FM allocations protection criteria. The report concluded "...protection criteria currently in use by the FCC provides at the so-called "protected" contour a signal which is listenable but of poor quality".

2. Subjective Evaluation of Audio Degraded by Noise and Undesired FM Signals (filed November 7, 1982). This report studied public subjective perception of noise-degraded audio signals. It found that protection ratios providing a stereophonic signal-to-noise (S/N) audio ratio of 50 dB are strongly preferred to those which yield only 30 dB.

3. Potential Effects of BC Docket No. 80-90 on FM SCA Operations (filed November 17, 1982). This report concluded that 80-90, as proposed, would have a negligible effect on stations employing subsidiary communications authorizations (SCA's) if the stations are operating on frequencies co-channel or adjacent channel to other stations. Second and third adjacent channel interference, however, "should present much more of a potential problem".

4. A Study of Stereo Radio Listening (filed December 7, 1982). "This study clearly indicates that stereo is the preferred method for listening to FM radio".

5. A Study of FM Station Mileage Separation (filed November 17, 1982). This report studied the distribution of actual mileage separations among domestic FM stations. The study concluded "that far more than the majority of FM broadcast stations provide a substantially better service at and within the 60 dBu contour than would be possible with a desired-to-undesired RF protection ratio of only 20 dB".

Analysis of Technical Studies and Relevance to Docket 80-90

The technical studies primarily address technical issues relating to one element of Docket 80-90's proposals: the desirability and impact of reduced required minimum mileage separations among FM broadcast stations, theoretically resulting in uniform protection from interference to FM broadcast stations within their 60 dBu (1 mV/m) contour. The results of these studies clearly imply that the separations among FM broadcast stations must not be reduced. To do so would cause unnecessary interference; destroy existing service that the public has become accustomed to; degrade second and third adjacent channel SCA operations; and afford insufficient protection to available FM "fringe" service.

But mileage separations are in essence just an administrative tool the agency uses to allocate FM stations. The engineering basis for their calculation lies with so-called "protection ratios"; the ratio of radio-frequency (RF) desired signal to undesired interference. When a given protection ratio is less than a certain standard value, interference is said to occur. In the co-channel case, for example, a stations's service is afforded "protection" if, according to present FCC Rules, the ratio of desired-to-undesired signal is at least 10 to 1.

The development of appropriate mileage separation standards is premised on the Commission's assumptions of FM service and interference that are embodied within its concept of protection ratios. Thus, to evaluate the suitability of the 80-90 proposed mileage separations, one must look to the definitions and assumptions of FM service and interference. Studies which engineers have viewed as essential to thorough examination of these assumptions include (1) assessing the capability of present and potential families of receivers to provide quality FM sound and (2) researching the value of audio signal-to-noise (S/N) ratio that constitutes "quality" and therefore warrants protection. Additionally, because there is a substantial "noise penalty" incurred by the operation of stereo broadcasting, the appropriate service to be protected -- whether mono or stereo -- must be determined.

The technical studies submitted in Docket 80-90 address these issues in substance. First, the listening study (#4) demonstrates that the public clearly prefers listening to radio in stereo and, therefore, <u>stereo</u> rather than mono is the service which should be protected. Second, the noise study (#2) demonstrates that acceptable quality of stereophonic service occurs upon provision of a 50 dB S/N ratio -- 316 to 1 -- at the protected contour -- 100 times better than existing FCC assumptions and a verification of the conclusions of the International Radio Consultative Committee (CCIR). Finally, according to the National Radio Systems Committee (the "NRSC" is a joint committee sponsored by NAB and the Electronic Industries Association (EIA) which evaluates transmission-reception system technical issues), a 50 dB stereophonic S/N ratio is achievable for many receivers if the RF desired-to-undesired ratios are engineered to <u>provide greater protection to FM service rather than less</u>. The NRSC recommended a co-channel protection ratio of 40 dB -- 100 to 1 -- a figure ten times greater than what the Commission presently uses. In other words, <u>existing minimum mileage separations provide for inferior FM stereophonic service and should be increased, or at least not decreased, if FM broadcasting is to continue, or improve, its high-quality transmission service to the public</u>.

If an inferior service is presently being provided, why, then, hasn't the Commission received large numbers of complaints? And why is FM generally regarded as the "high-quality" aural medium?

There are two principal answers to these questions. The study of mileage separations (#5) shows that actual mileage separations among FM broadcasters are, on average, substantially larger than the minimum required separations. For example, the average distance to the three nearest co-channel neighbors of Class A stations has a median value of 105 miles. By way of contrast, the existing specified minimum separation is 65 miles and Docket 80-90 proposes to reduce this value to 61 miles. For a protected contour of 60 dBu (1 mV/m), a separation of 105 miles corresponds to a co-channel protection ratio of 35 dB -- 56 to 1 -- assuming that both stations are employing maximum power/height. Thus, Class A stations are receiving, on average, protection that is within 5 dB of the values originally specified in NRSC's report as necessary to provide a quality FM radio service: a 50 dB stereophonic S/N. For Class B and C stations, a similar situation exists: actual co-channel protection, on average, is 40 dB and 34 dB, respectively.

It is clear, then, that in general the service of FM broadcast stations enjoys generous protection not as the result of specific FCC policies, but because the juxtaposition of FM assignments is such that median separations among stations are larger than FCC-required minimums.

Second, when interference does occur -- i.e. at minimum separations -- it is not recognized as such by the general public and is thus not immediately attributable to mistaken FCC allocation policies. Instead of hearing the program of the undesired stations, as in AM or "CB" type interference, FM interference is similar to commonplace "ignition-system" noise -- a staccato series of clicks and pops. This effect is dissimilar to AM or CB interference which the general public is more likely to perceive as interference to a desired broadcast from another station.

But the result of FM interference is identical to the result of any form of interference: an otherwise desired signal is precluded from adequate reception.

Conclusion

For the engineering reasons stated herein, any proposed reduction in the minimum required mileage separations among FM broadcast stations disserves the overall interests of broadcasters. While actual median separations provide the listener with good quality service, interference unavoidably occurs and FM service is needlessly degraded where these separations are at a minimum. Accordingly, any further reduction of minimum separations only serves to increase aggregate levels of interference.

Michael C. Rau

National Association of Broadcasters
March, 1983

Report on SCA Operation

Robert W. Denny, Jr.

Jefferson-Pilot Broadcasting Company

Charlotte, North Carolina

SCA or Subsidiary Communications Authorization as applied to FM broadcasting is the simultaneous transmission of a program in addition to the regular monophonic or stereophonic program. SCA programming is highly specialized and not of interest to the general public. Some examples of SCA programming are background music for businesses, informational services for professional groups, "talking book" services for the visually handicapped, and the control of power demands. An SCA transmission need not be limited to aural programming. SCA may transmit any data encoded using multifrequency signalling techniques, such as slow-scan TV pictures and teletype. The SCA multiplex transmission may itself be a frequency multiplex signal, comprised of several unique data and aural channels. This process would allow multiple aural and data signals to be combined at the SCA injection point and separated at the user's receiver.

With the increasing fragmentation of radio station listenership, the marketplace for the sale of advertising is becoming more competitive. This fragmentation, together with inflation, is causing the expenses of operating a radio station to increase faster than station income. At the same time, the demand for SCA distribution of information is increasing at an unprecedented rate.

Small market licensees have been aware of SCA's potential for producing revenue for years. Until now, in the larger, more profitable operations, additional revenue produced through the operation of an SCA was not significant when compared to revenues produced from the sale of time on the main stereo channel. This fact, coupled with the inherent degradation of the main channel stereo programming due to SCA operation, has resulted in limited acceptance of SCA by owners, general managers, program directors, and engineers.

The purpose of this paper is to examine the effect of SCA operation on the main channel stereo programming and to establish a set of in-house standards for SCA operation. These standards should allow the operation of an SCA subcarrier with minimum degradation to the main channel. By examining the degradation resulting from SCA operation and offering solutions to eliminate or minimize it, this paper asserts that broadcasters can enjoy the best of both worlds: The additional revenue procured from SCA operation, and the highest quality operation of the main channel attainable.

There are two causes of degradation to the stereo program, particularly the double sideband surpressed carrier L-R signal, when a 67 kHz SCA multiplex signal is added to a stereo baseband. These are:

1. Beat notes
2. Crosstalk

Beat notes can result from two conditions. They are:

1. Diode Matrix Decoder
2. Multipath Propagation

BEAT NOTES

Diode Matrix Decoder

A 9 kHz beat note falling in the L+R main channel (mono) portion of the baseband signal is generated in the receiver's stereo decoder. It can sound like a high pitched whine to the stereo listener. This beat note is produced when the 67 kHz subcarrier (SCA) and the second harmonic of the 38 kHz subcarrier (stereo) are combined.

There are two types of stereo decoders used in the FM stereo receivers that are in the listeners' hands today. One is the diode matrix decoder, which is giving way to the newer phase lock loop (PLL) decoder. John Keane of National Public Radio found, "When the (SCA) trap is used (in the diode matrix type stereo decoder), the 9 kHz beat note is reduced..., for a net level of -60 dB referred to 82% (peak stereo) modulation. In contrast, the 9 kHz beat note level from the matched output of the PLL decoder (RCA CA1310) is -71 dB."[1] The performance of the older receivers is marginal for good stereo reception. Since the superior PLL decoders dominate the receiver marketplace and will, in time, replace the diode matrix decoders, the 9 kHz beat note will cease to be a problem. The PLL stereo decoder is largely responsible for the newly found SCA acceptance by broadcasters.

Multipath Propagation

Second order intermodulation distortion also produces a beat note. Mitsuo Ohara of NHK subjectively determined that -55 dB is the level above which this beat note becomes perceptible.[2] This beat note differs from the previously discussed beat note in two ways: 1) It falls at 10 kHz in the L+R main channel (mono) portion of the baseband signal; and 2) it is not generated by the stereo decoder in the listener's receiver.

The 10 kHz beat note is generated by the combination of the 19 kHz stereo pilot carrier and the 67 kHz SCA subcarrier in the presence of multipath propagation. As multipath propagation is regularly found in the vicinity of large surfaces capable of reflecting VHF signals, such as mountains or tall buildings, the effect of this beat note is certain to be noted in some areas. It is very difficult to predict the areas in which this 10 kHz beat note degradation will occur. This is because the occurrence of the condition is not only a function of the ratio of direct to reflected signal intensity at the receiver, but one of the delay between the direct and reflected signals' arrival at the receiver as well. At a delay time of 15 microseconds, the ratio of direct to reflected signal intensity need only be about three to overcome this beat note. At a delay time of 25 microseconds, this ratio must be over 30.[3]

This mode of degradation is the most significant because it is highly unpredictable, uncontrollable, and sure to occur on some receivers in certain locations. The only relief is the use of a directional antenna, an item not usually found in cars. Possibly the only way to discount the effects of multipath induced IM and the 10 kHz beat note is to consider that multipath most severely affects the sereo signal itself. That is, in the presence of multipath, the stereo signal is already distorted, and the addition of the 10 kHz beat note will only make the stereo program more distorted.

CROSSTALK

Crosstalk from the SCA channel to the stereo subchannel (L-R DSBSC) can be caused by two conditions. These are:

1. Multipath propagation
2. Modulation Index

Since the uncontrollability of multipath propagation was discussed above, no further suggestions will be made for solving this problem. In areas of high multipath distortion, the stereo signal will probably be unacceptable to the listener, and the 10 kHz beat note effects can be discounted. However, the effects of multipath on the reception of the stereo program in marginal areas should be mentioned. To a listener in an area of light multipath interference, which he may tolerate, the modulated SCA will be heard as "birdies," what Ohara calls "monkey chatter." The addition of this noise, which is merely the modulated 10 kHz beat note with the associated sidebands, to the already present multipath distortion, may create a condition which will cause the once tolerant listener to change stations.

Modulation Index

The final cause of degradation, overdeviation and the transmission of non-bandlimited audio on the SCA, is possibly the easiest to control and the most overlooked. If the transmission facility is of reasonable linearity, and these two situations are ignored, degradation will occur in the form of "birdies" and swishy noises in the stereo program.

If the SCA subcarrier is overdeviated (overmodulated), or the modulation is not severely band limited, crosstalk will result. This crosstalk results when the SCA's lower sidebands overlap with the upper sidebands of the double sideband supressed carrier stereo difference channel (L-R). When this overlapping occurs and is decoded by a stereo decoder, similar swishing noise and "birdies" are heard, even if a PLL stereo decoder is being used, and conditions for reception are ideal.

The broadcaster can control this degradation, because it is produced in the exciter, far before the transmitter or antenna. Allowing these conditions to persist results in a transmitted signal outside the parameters defined by the

Federal Communications Commission for SCA operation.

FCC Requirements vs Lessee Requirements

¶73.319(e) of the Commission's Rules and Regulations states, "Frequency modulation of the main carrier caused by the SCA subcarrier shall, ... in the frequency range 50 to 53,000 Hz, be at least 60 dB below 100 percent modulation." If a 67 kHz subcarrier is injected at 10%, or 20 dB below the main carrier, the third order lower sideband component of a subcarrier modulating tone, falling in the region 50 to 53,000 Hz, must be down to an amplitude of 1%, or 40 dB below the unmodulated 67 kHz subcarrier. This corresponds to a level 60 dB below the main carrier. Lower sideband analysis can easily be handled through the use of Bessel Functions. By knowing the SCA injection level (which is 10% unless noted otherwise); maximum peak deviation, $\Delta f_{c\ max}$; and maximum modulating frequency, ρ_{max}; one can make a complete analysis of all subcarrier sidebands.

Since most SCA subcarriers are leased to a third party by the station, it is not unusual for this third party to specify the technical requirements of the SCA channel to be delivered by the station. Presented here as an example are the specifications for an SCA channel from a contract recently considered by Jefferson-Pilot Broadcasting Company.

Subcarrier Frquency $\equiv f_c$ = 67 kHz, ±500 Hz

Injection Level = 10%

Peak Deviation $\equiv \Delta f_{c\ max}$ = ±5 kHz

Max. Modulating Frequency $\equiv \rho_{max}$ = 5 kHz

To test for compliance with 73.319 (e), the modulation index is calculated using the maximum values:

$$\text{Modulation Index} \equiv m = \frac{f_{c\ max}}{\rho\ max} \quad \frac{5\text{ kHz}}{5\text{ kHz}} = 1$$

The SCA subcarrier lower sidebands, which pose a potential interference problem with the stereo L-R DSBSC signal, will recur at each multiple of ρ_{max}:

Sideband Order, n	f (kHz)
Carrier	67
1	62
2	57
3	52

The amplitude of these lower sidebands, as a percentage of the unmodulated 67 kHz subcarrier, can be determined through the use of Bessel Funtions of the form $J_n(m)$ for m = 1:

Sideband Order, n	f (kHz)	Bessel Value $J_n(m)$	Amplitude as a Percentage of Subcarrier	Amplitude (dB)
0	67	0.7652	76.52	-22.32
1	62	0.4401	44.01	-27.13
2	57	0.1149	11.49	-38.79
3	52	0.0196	1.96	-54.15

As shown, the third order lower sideband (n=3) falls at 52 kHz, which is well inside the L-R DSBSC spectra. For this modulation index, m=1, and the specifications given by the lessee, the third order lower sideband is only 52.15 dB below the level of the main carrier. Not only would SCA operation under this set of conditions not be in compliance with ¶73.319 (e), but the station's stereo main channel program would be degraded in a way perceptible to the listener. Surprisingly a number of stations, have accepted these specifications.

A Closer Look at ρ and Δf_c

Before accepting specifications regarding ρ_{max} and $\Delta f_{c\ max}$, it is important to note that for $\rho \geq 3{,}500$ Hz, these parameters cannot be selected independently. This is because the plot of ρ vs. Δf_c that produces compliance with ¶73.319 (e) is not a smooth function. There are discontinuities at 3,500 Hz, 4,667 Hz, 7,000 Hz, and 14,000 Hz. At these values, a new order lower sideband enters the region of the baseband 50 to 53,000 Hz. The maximum peak deviation at these critical frequencies is shown in Table I. A graph showing ρ_{max} vs. $\Delta f_{c\ max}$ in compliance with ¶73.319(e) appears in Figure I.

TABLE I

Points of Discontinuity

ρ (Hz)	$\Delta f_{c\ max}$ (Hz)	$J_n(m) = 0.0100$
3,500	±5,022	$J_4(1.4350)$
4,667	±3,699	$J_3(0.7925)$
7,000	±1,992	$J_2(0.2845)$
14,000	± 280	$J_1(0.0200)$

Inspection of Table I and Figure I reveals Δf_c can be ±5 kHz as long as $\rho = 4{,}667$ Hz. At $\rho = 4{,}667$ Hz, $\Delta f_{c\ max}$ must decrease to ±3.7 kHz in order to comply with ¶73.319(e). From $\rho = 4{,}667$ Hz to $\rho = 5{,}000$ Hz, $\Delta f_{c\ max}$ can increase from ±3.7 kHz to ±4 kHz as shown in Figure I.

A VIABLE SYSTEM

Keeping in mind the necessities of controlling modulation audio bandwidth and SCA subcarrier deviation, a prototype system was assembled and tested on the air from January 4, 1982 to April 1, 1982, under a temporary authority granted by the FCC. This system consisted of a Broadcast Electronics, Inc. FC-30 SCA

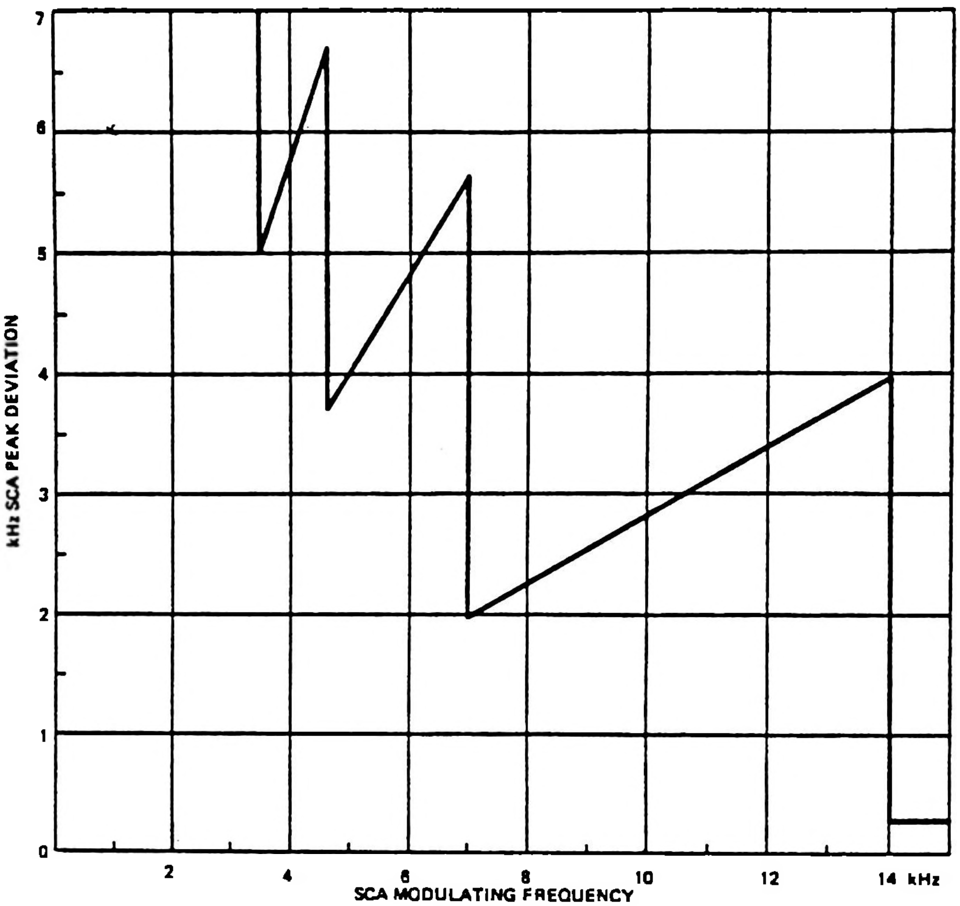


Figure 1, Points of Discontinuity

Maximum deviation of 67 kHz, 10% amplitude SCA Subcarrier with stereo broadcasting [4] generator was selected because the unit's audio input conditioning filter most closely approximated the ideal filter, both in cut-off frequency and number of poles. Also, the generator was designed for use with the FX-30 FM exciter used by the station.

System Configuration

The equipment was disposed as shown in Figure 2. For the tests I will describe, the SCA subcarrier was injected at a port of the PCL-505 studio transmitter link provided for this purpose. In an actual system, I strongly recommend that the SCA program be delivered to the transmitter via land line, and the SCA subcarrier injected directly into the FM exciter, to eliminate any distortion of the FM stereo signal, produced by the STL modulator demodulator.

Crosstalk Testing

Tests of the crosstalk from the SCA subchannel into the L-R subchannel were made using a TFT 724A modulation monitor and a Tektroniks 5L4N spectrum analyzer. Figure 3 shows the 67 kHz subcarrier modulated 100% with 3kHz. At 53kHz, the lower sidebands resulting from the SCA operation are -62dB, referenced to the main carrier. This would indicate compliance with 73.319(e) and minimal degradation of the L-R signal. Figure 4 shows the same 67 kHz subcarrier modulated 83% as indicated on the modulation monitor with 4.7 kHz. The third order lower sideband, falling at 52.9 kHz, is down only 48dB, with respect to the main carrier. By reducing the modulation level to 61% with the audio input conditioning filter, the third order lower sideband is eliminated, as shown in Figure 5.

CONCLUSION

With the dominance of phase lock loop FM stereo receivers in the consumer marketplace, the receiver generated distortion which once plagued SCA operations is rapidly disappearing. If a broadcaster is willing to accept further degradation of the FM stereo signal in multipath areas, which may already be unlistenable, and to sacrafice 10% of the station's loudness to the SCA subchannel, SCA can become an additional source of revenue for the station. I should stress, however, that SCA operation without degradation of the FM stereo program depends upon the following:

1. Linear, state of the art FM transmission system
2. Good transmission system maintenance
3. Regular monitoring of the complex FM waveform with a spectrum analyzer, to assure compliance with 73.319 (e)
4. Limited audio bandwidth and SCA generation system

Adherence to these four points will increase operating expenses, as long as the SCA is in operation. There will also be some start-up expenses, to purchase equipment and to revamp the transmission system , if needed. I believe that these costs, both long and short term, will be quickly recoverd by the long term lease of the SCA subcarrier.

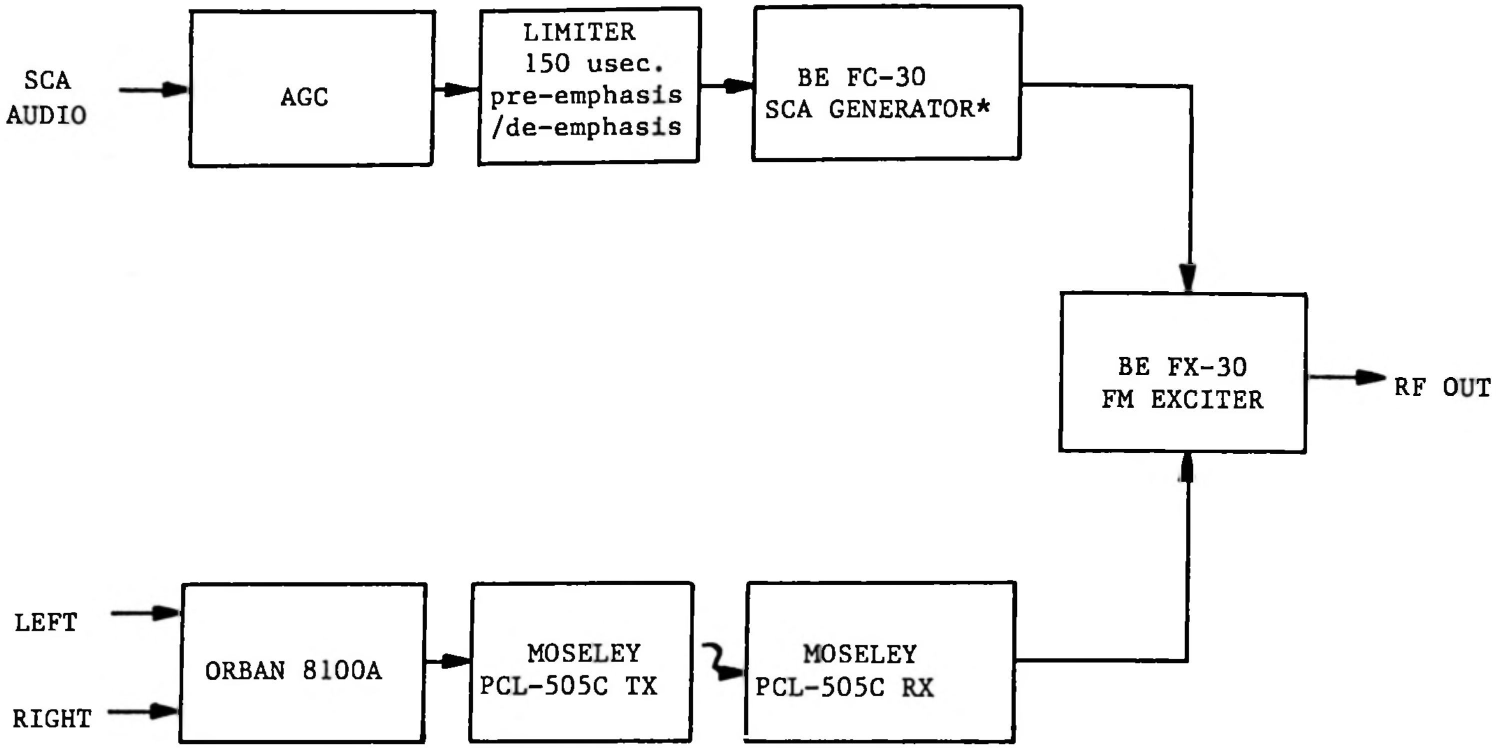


Figure 2, Equipment Configuration

* Broadcast Electronics FC-30 SCA Generator includes audio input conditioning filter which consists of an active 6-pole (-36 dB/octave) low pass filter, -3 dB at 4.3 kHz.

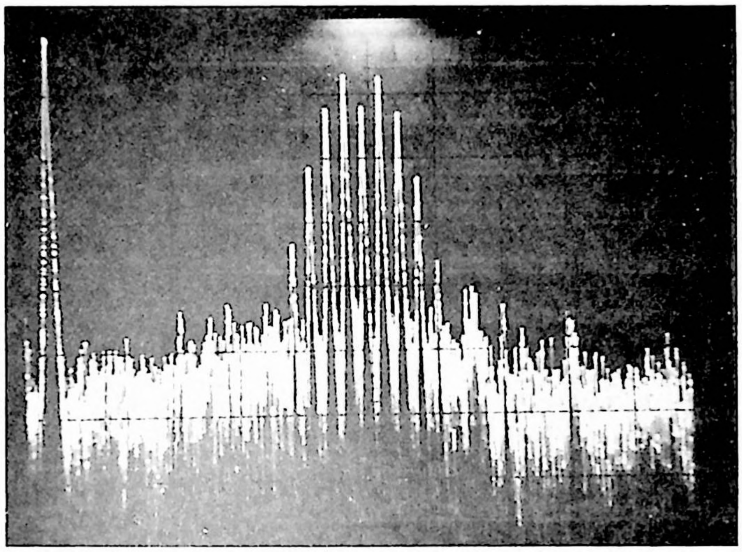


Figure 3

67 kHz subcarrier, 10% injection, modulated 100% with 3 kHz, 5 kHz deviation.

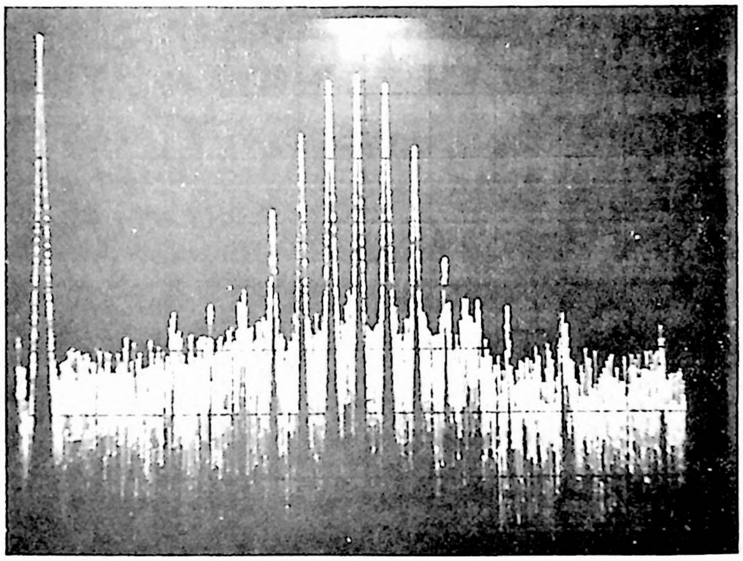


Figure 4

67 kHz subcarrier, 10% injection, modulated 83% with 4.7 kHz, 100% modulation is 5 kHz deviation. Note third order LSB in L-R subchannel at -48 dB level with respect to carrier.

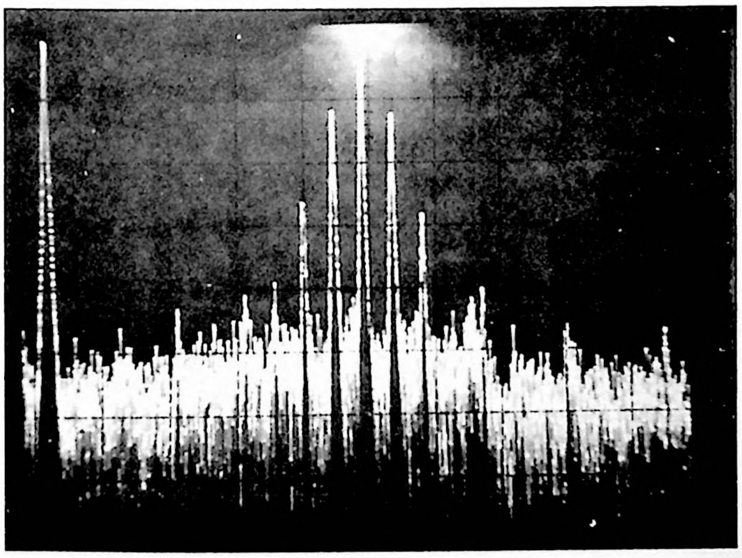


Figure 5

67 kHz subcarrier, 10% injection, modulated 61% with 4.7 kHz, 100% modulation is 5 kHz deviation. Note third order LSB in L-R subchannel is down to -60 dB level with respect to carrier.

References

(1) John Kean, "Laboratory and Field Tests of Several Fm/SCA Frequencies," National Public Radio Engineering Update, Vol. II, No. 7, p. 12 (1981 Oct.).

(2) Mitsuo Ohara, "Distortion and Crosstalk Caused by Multipath Propagation in Frequency-Modulation Sound Broadcasting," IEEE Transactions on Broadcasting, Bol. BC-26, No. 3, p.81 (1980 Sept.).

(3) Ohara, p. 80.

(4) Warren B. Bruene, "FM Broadcast Transmitters," in Engineering Handbook, ed. George W. Bartlett (Washington, D.C.: National Association of Broadcasters, 1975, p. 445.

THE EFFECT OF ADDITIONAL SCA SUBCARRIERS ON FM STEREO PERFORMANCE AND RF PROTECTION RATIOS

John Kean, Senior Engineer,

National Public Radio,

Washington, D.C.

This is a summary of the results of RF protection ratio tests conducted with SCA subcarrier frequencies and peak carrier deviations proposed in the FCC's Notice of Proposed Rule-Making 82-536. Comments on the preliminary findings are included.

Technical comments on rule making 82-536 were generally very favorable regarding the Commission's proposal to extend the FM baseband to 99 kHz and to allow an increase in total peak modulation when two or more SCAs are transmitted.*

However, two commenters urged more tests to confirm that no ill effects would result from the proposed changes.
From the National Association of Broadcasters:

> "NAB suggests that the Commission undertake studies, including ample testing, to determine precisely what the effect on the technical integrity of the FM broadcast system would be if the Commission were to authorize a higher total modulation percentage than presently allowed and/or higher baseband frequencies."

And from the Harris Corporation:

> "The Commission may want to conduct a thorough engineering study as to the impact of the two proposed technical rule changes involving the increase in maximum modulation deviation and the increase in SCA sidebands."

NPR believes that concern of interference from added SCA operations falls into three specific areas, to be discussed in this paper:

---

* In its comments to the Commission in 82-536, NPR urged that the FM modulation rules be changed to permit a main-channel backoff equal one-half the total SCA injection; this may result in a total peak deviation of up to 110% with two SCA sucarriers.

- o Interference to the FM carrier's main channel service;
- o Interference to an existing SCA service on the same FM carrier;
- o Interference to co-channel, and first- or second-adjacent channel FM stations.

INTERFERENCE TO THE FM CARRIER'S MAIN CHANNEL SERVICE

Degradation could result from several causes, which have been adressed by NPR in previous studies. A prime source of this interference is the chance of audible beat notes, caused by second order IM distortion in the overall broadcast/receiving system. While distortion caused at the transmission end is minimal in a properly-operating transmitter and antenna system, difficulties at the receiving end, such as multipath and receiver mis-tuning, are possible. Fortunately, the cumulative effect of these problems, both for single and multiple SCAs, is tolerably small. The effects are examined in the earlier NPR technical paper "Laboratory and Field Tests of Several FM/SCA Frequencies," which follows this report.

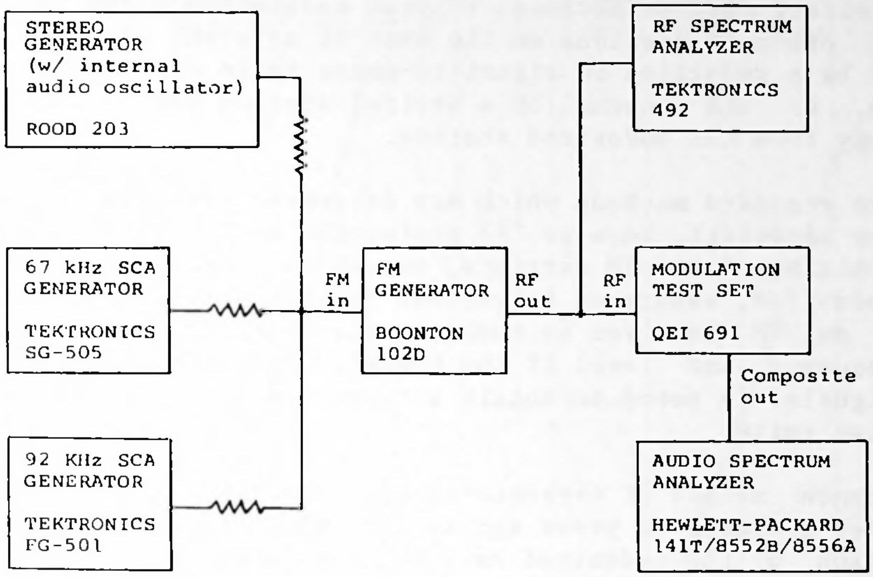
Another source of possible degradation under the proposed SCA rules is distortion resulting from increased peak deviation. Tests conducted by NPR in the laboratory and over-the-air tests indicate no measureable effect on main channel monophonic or stereophonic performance due to the increased peak deviation. This is because of the very low RMS modulation increase (and corresponding RMS bandwidth contribution) of SCA subcarriers, and the low statistical probability that all signals modulating the carrier reach their maximas at the same instant.

INTERFERENCE TO AN EXISTING SCA SERVICE ON THE SAME STATION

The above condition holds for main channel-to-SCA crosstalk, despite the "increased" modulation in the proposed rules. In fact, the main channel modulation would be the same with an additional subcarrier as it would be under present rules with a single subcarrier. The modulation added by a second SCA (having a modulation index of 7.5/92, or 0.08) has a very small effect on an existing subcarrier. Since loss of main channel modulation potential, i.e. loudness, is a major concern to FM station management, this policy aids acceptance of SCAs by broadcasters while increasing the market supply.

NPR has conducted tests to evaluate occupied bandwidth under various modulation conditions, using the equipment shown in Figure 1. The change in sideband energy from a stereo-only condition to a stereo-plus-two SCA condition is graphically shown in Figures 2 and 3. Within 150 kHz of the carrier there is no increase in sideband energy. An increase can be found only for sidebands beyond 150 kHz and at levels more than 50 dB below the FM carrier.

Figure 1 - FM MODULATION TEST EQUIPMENT



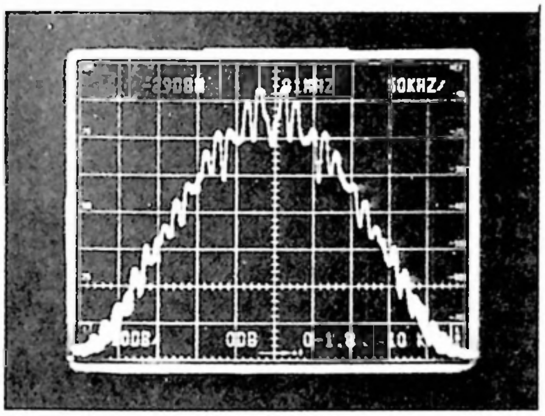


figure 2

Overlay of RF Spectra

View A: 91% mod. @ 15 kHz L. ch. (stereo) 9% Pilot

View B: 82% mod. @ 15 kHz L. ch. (stereo) 9% Pilot 9% 92 kHz SCA

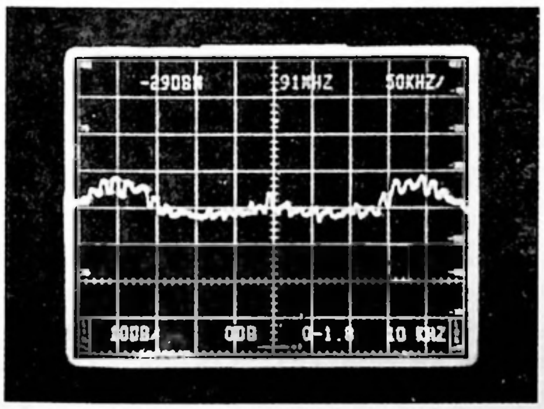


figure 3

Difference in RF Spectra

View A: same as fig. 4
View B: same as fig. 4

Note: net decrease in sideband energy within ±150 kHz of carrier, increase beyond for very low level sidebands

INTERFERENCE TO CO-CHANNEL, AND FIRST- OR SECOND-ADJACENT CHANNEL FM STATIONS

As far as an increase in interference, NPR believes the chief concern lies in the effect that an increase in peak modulation might have on the reception of other FM stations on the same or adjacent channels. The effect would be a reduction in signal-to-noise ratio of the demodulated mono, stereo, or SCA channel of a desired station due to increased sideband energy from the undesired station.

There are standard methods which may determine the effects of changes in modulation bandwidth, know as "RF protection ratio tests." The tests involve the mixing of two FM carriers, one which is desired and a second which is undesired, separated by various frequency spacings and level differences: an FM receiver is tuned to the desired carrier, and the relative frequency and level of the undesired carrier (with various modulating signals) is noted to obtain a reference signal-to-noise ratio.

The common method of determining RF protection ratio in the U.S. was established a number of years ago by the IEEE. It uses full monophonic tone modulation of the undesired carrier, and seeks a protection ratio which yields a 30 dB quasi-RMS signal-to-noise ratio of the demodulated, desired carrier.

A more recent method, adopted by the International Electrotechnical Commission and others, uses weighted noise at 32 kHz quasi-peak deviation on one stereophonic channel. This signal is said to better represent the audio spectrum and modulation characteristics of typical program material. It seeks a protection ratio which yields a 50 dB quasi-peak, mid-band frequency-weighted, signal-to-noise ratio of a demodulated, stereo-decoded desired carrier.

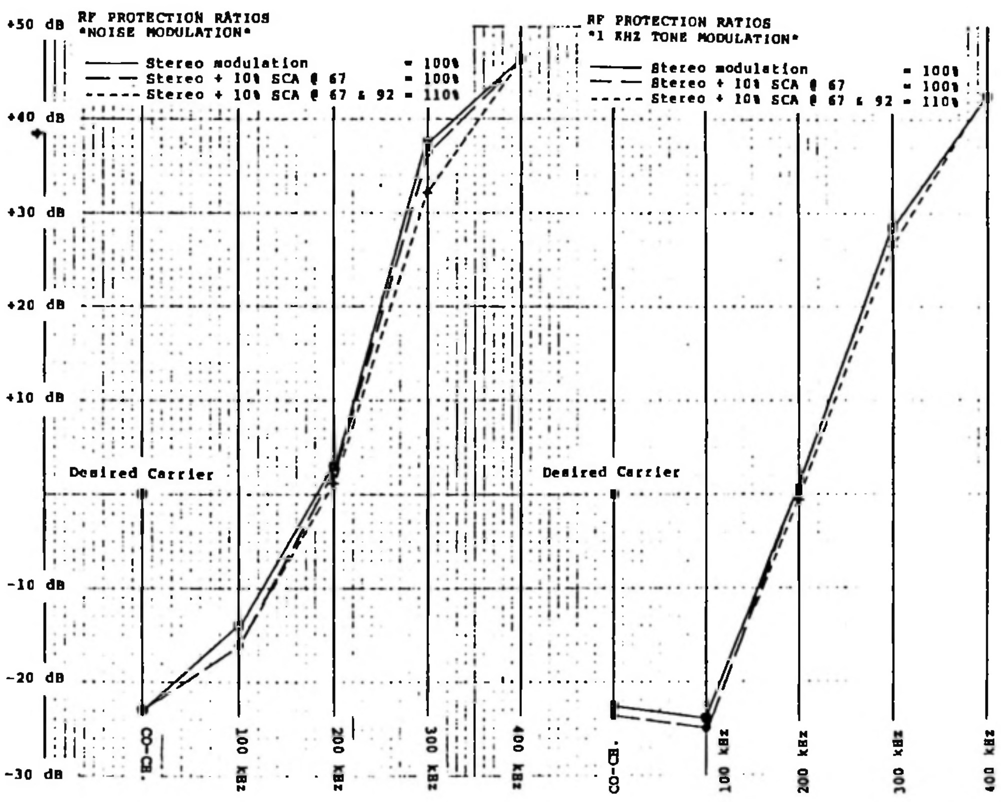
The following tests were conducted using variations of both methods, as noted.

PROTECTION RATIO MEASUREMENTS
ADJACENT CHANNEL (+/-200 KHZ)
DESIRED CARRIER IN STEREO

Undesired Mode:	Pink Noise @ -12 dB	Full 1 kHz L+R
Monophonic	-1.5 dB	- db
Stereophonic	0	0
Stereo + SCA @ 67	-1.0	+0.4
Stereo + SCA @ 92	-1.7	-0.4
Stereo + SCA @ 67 & 92	-2.2	-0.6

Note: -Measurements from Left channel of high-quality FM tuner (Phillips AH-673) for 50 dB stereo signal-to-noise ratio
-De-emphasized noise measured true-R.M.S., unweighted
-Pink noise was band-limited, pre-emphasized, with approximate 6 dB crest factor (peak-to-RMS ratio)
-Values are average for upper and lower adj. channels
-Normalized for stereo-to-stereo condition

Figure 4



A 50 dB noise reference was chosen because it better represents minimum performance for high-fidelity channel than does 30 dB. A quasi-peak reading psophometer was not available for the measurements, however, the audible effect of the noise was free of "clicks" and other coloration. The pink noise modulation did not conform to IEC 315-4, however, the signal was 15 kHz bandlimited and pre-emphasized, and was adjusted to an approximate 32 kHz quasi-peak modulation. This resulted in a -12 dB RMS modulation, referred to 100% tone. Tests were conducted co-, first-, and second-adjacent channel. In the +/-400 kHz case, there was either no measureable effect or the differences were too small and random to be meaningful.

Figure 4 shows in graph form the RF protection ratios for the stereo-only and the stereo-plus-dual SCA mode (with 110% peak modulation). Both upper and lower side measurements are averaged in single curves. For the reasons mentioned above, the absolute values measured may differ from the IEC standard tests, (but the tabulated ratios between frequencies of the same mode should be valid).

The conclusion which may be drawn from this basic test is that the proposed modulation rule changes with SCA would have a small-to-negligible effect on RF protection ratio, depending on how one chooses to represent main-channel modulation. I assume that the IEC (and the CCIR) chose 32 kHz modulation because it represented the long-term average of lightly processed (peak-limited) program audio.

I agree that a 50 dB stereo noise degredation, possibly psophometrically-measured, is better at matching the ear's sense of audio channel quietness, and is certainly superior to the old 30 dB monophonic reference. However, for the purposes of measuring the true effect of a change in modulation rules on U.S. FM stations, I doubt that low-level pink noise represents current practice: even with stations which claim minimal audio processing, it is not uncommon to see RMS modulation of 70% or more on loud, sustained passages.

More tests would be helpful in gaining a picture of the practical effect of the proposed modulation rules as SCAs are added. For example, the effects to desired SCAs on stations with first- or second-adjacent-channel spacing would provide useful data (although SCAs do not enjoy a protected status as does main channel program service). NPR intends to continue testing and analysis in order to advance and further document the state of the art in FM subcarrier transmission.

LABORATORY & FIELD TESTS OF SEVERAL FM/SCA FREQUENCIES

In the summer of 1981 a series of laboratory and field tests were conducted to evaluate several FM/SCA frequencies which are above the range presently authorized by the FCC. It is believed that one of these frequencies would permit the addition of a second SCA service on a stereo FM station, 1/ or would reduce potential interference to the station's stereo service. 2/ The tests were designed to try these assertions and document their performance relative to the 67 kHz subcarrier.

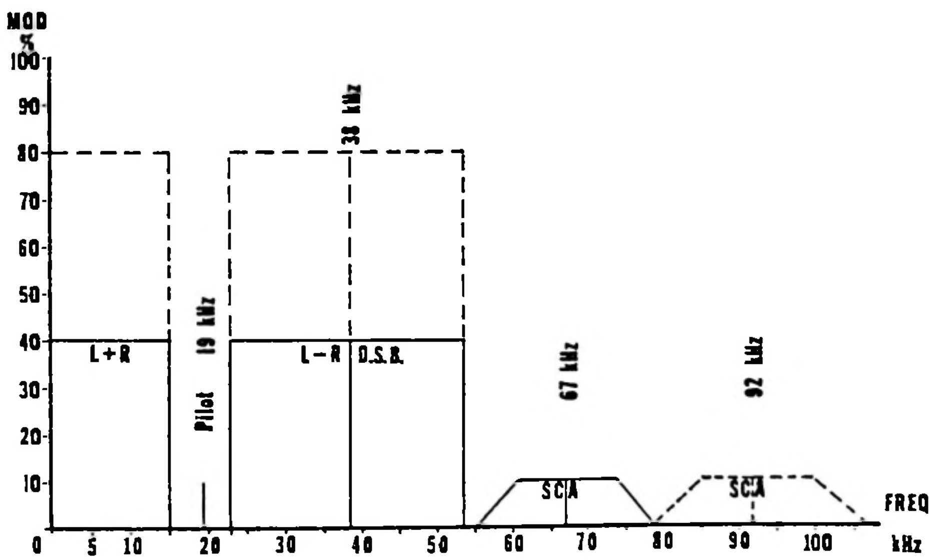
The basic findings are that 92 kHz is the best choice for a new aural SCA service; that its performance is similar to 67 kHz SCA; that it produces lower interference levels to main channel stereo service than 67 kHz, and that it can be successfully operated in addition to stereo and existing SCA services.

BACKGROUND OF PRESENT FM MULTIPLEX

Introduction of new signals into the modulating baseband of an FM station requires attention to its possible interactions with existing baseband signals, as well as its chosen performance characteristics. The familiar baseband spectrum of the combined stereo and SCA signals is shown in Figure 1.

Figure 1

FM Stereo and Subcarrier Baseband



The modulating frequency is shown on the horizontal axis; the modulation level as a percentage of 75 kHz peak deviation is shown on the vertical axis. In the main channel (direct modulation by the L+R signal), the maximum deviation may not exceed 80 percent of the total deviation, assuming identical amplitude and phase on the L and R signals. Otherwise, the modulating signal is divided between the L+R and L-R channels. The Pilot signal at 19 kHz deviates the main channel within the limits of 8 to 10 percent. The SCA subcarrier at 67 kHz contributes up to 10 percent of the total modulation. A nominal value of 9 percent has been shown. The arithmetic sum of the three deviations approaches 100 percent which is ± 75 kHz deviation.

The 92 kHz subcarrier used in the field tests is shown in Figure 1 to be added to the existing baseband signals of WETA-FM, which provided the transmission facilities. While this could occasionally result in a peak deviation of 110 percent (40 + 10 + 40 + 10 + 10 = 110), it was felt that the RF bandwidth of the total FM carrier would not be significantly increased. It was found during the tests that neither the -3 dB nor -20 dB bandwidths were noticeably increased by this condition, since the 92 kHz subcarrier creates a pair of primary sidebands well below this level.

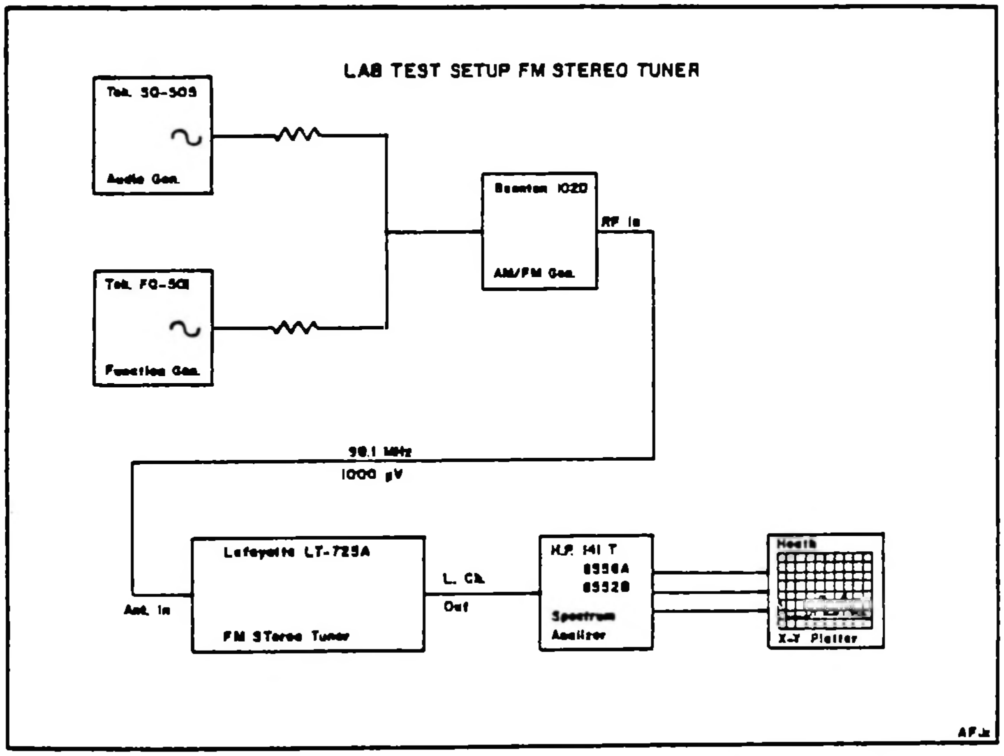
Adding the test subcarrier to the existing baseband signals had the advantages of comparing performance of both subcarriers with "normal" modulation (up to 10 percent injection for each) and avoided problems of shifting the main channel modulation up or down while maintaining stereo Pilot injection. It also created a slightly greater chance for interference to stereo listeners, which would make identification easier.

MECHANISMS OF SCA/STEREO DECODER INTERFERENCE

Figure 2 shows the test setup for a representative FM stereo tuner, using the following equipment:

- Tektronix SG-505 Audio Generator (19 kHz @ 7.5 kHz dev.)
- Tektronix FG-501 Function Generator (SCA @ 7.5 kHz dev.)
- Boonton 102D AM/FM Generator (98.1 mHz, 1000 uV out)
- Lafayette LT-725A FM Stereo Tuner
- Hewlett-Packard 141T Spectrum Analyzer System
- Heath X-Y Plotter

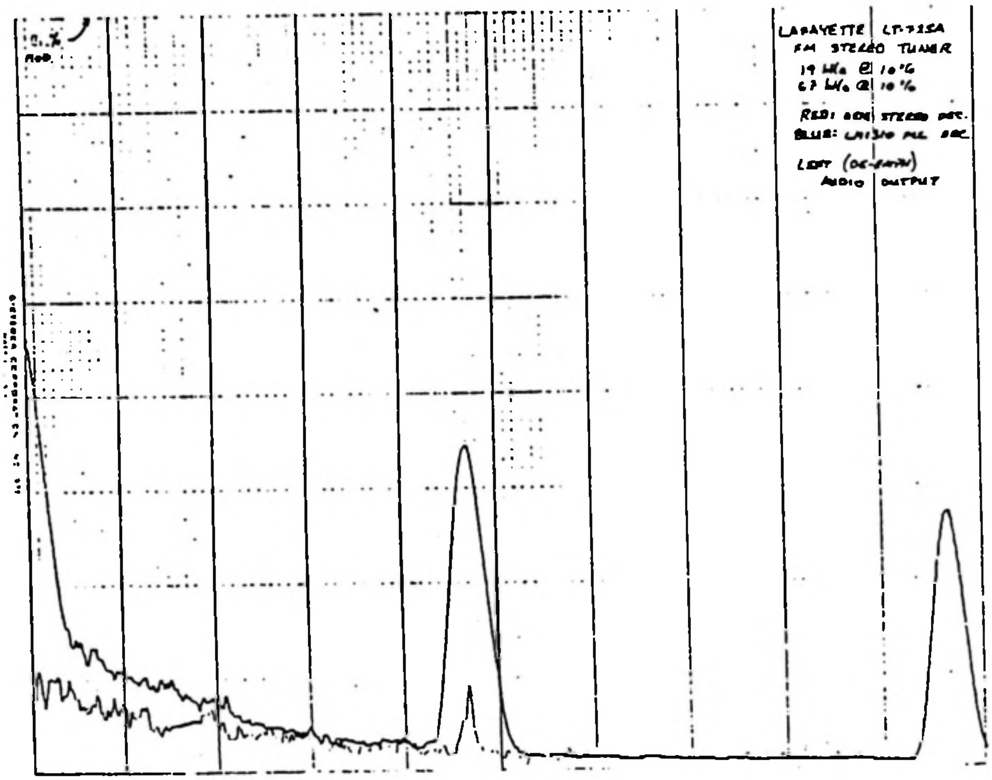
Figure 2



The spectrogram in Figure 3 shows the demodulated, de-emphasized Left channel of the sample tuner. This tuner was used because of its older design (it utilizes a four-stage IF with double-tuned transformers between each and a discrete-component stereo decoder having a diode-matrix type output), which was fairly common thru the early 1970's.

Two beat-note signals are shown at 9 kHz, the higher one produced by the original discrete-component stereo decoder, the lower produced by an RCA CA1310 Phase Lock Loop IC Stereo decoder installed in the tuner. Stereo Pilot leakage at 19 kHz is visible at the extreme right.

Figure 3



Horizontal Scale: 0-20 kHz, 2 kHz/div.,
Vertical Scale: 10 dB/Div.,
top line is 80% of Total Modulation.

It should be noted that the 67 kHz SCA trap was removed from the original stereo decoder for this and other spectrograms. This was done to permit direct comparisons of beat-note levels as a function of SCA frequency. When the trap is used, the 9 kHz beat-note is reduced 14 dB at the decoded stereo output, for a net level of -60 dB referred to 82% (peak stereo) modulation. By contrast, the 9 kHz beat-note level from the matched output of the PLL decoder is -71 dB. No other coherent components are visible or audible in either spectrogram.

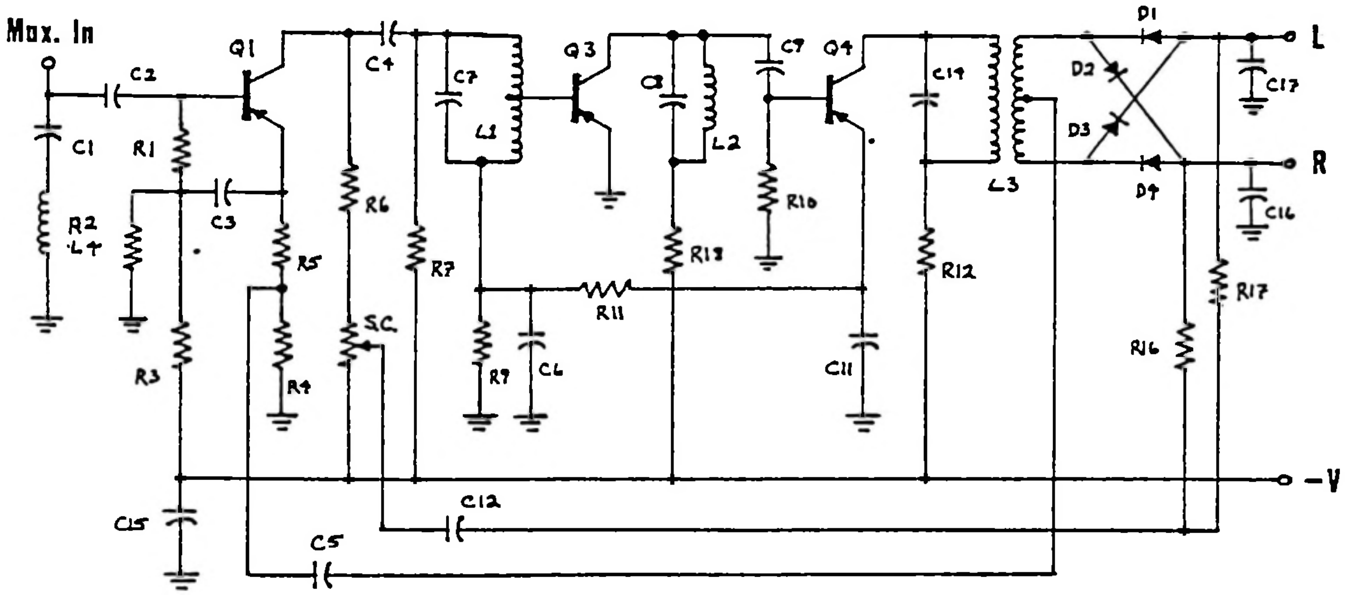
The 9 kHz beat-note is produced by the difference in frequency between 67 kHz (SCA subcarrier) and 76 kHz (twice the L-R subcarrier) as:

9 kHz = 76 kHz [38 kHz x 2] - 67 kHz

A review of the operation of the two stereo decoders will explain the difference in beat-note levels. Figure 4 shows the simplified schematic of a discrete-component stereo decoder similar to the original one in the sample tuner.

Figure 4

Discrete-Component Stereo Decoder, Simplified



The circuit comprises input amplifier Q1, providing separate baseband outputs of opposite phase at the junction of R4/R5 and S.C., the "Separation Control." Q3 is a 19 kHz Pilot amplifier, using C7/L1 and C8/L2 as tuned input and output filters. Q4 is operated as a driver for C14/L3, a 38 kHz resonant "flywheel" synchronized to the

19 kHz driving signal. The secondary of L3 provides a high-level 38 kHz sinewave to the ouput diode matrix. These diodes (D1/D3 and D2/D4) form a pair of switches with opposite phase outputs at C17 and C16, respectively.

When composite baseband audio is fed through C5 to the centertap of the L3 secondary, a synchronously-detected audio waveform appears across the C17 and C16, corresponding to the L-R signal modulating the original 38 kHz double-sideband supressed carrier. This signal is then combined with the L+R signal fed thru R16 and R17 to produce the total Left and Right channel signals from: (L+R) ± (L-R).

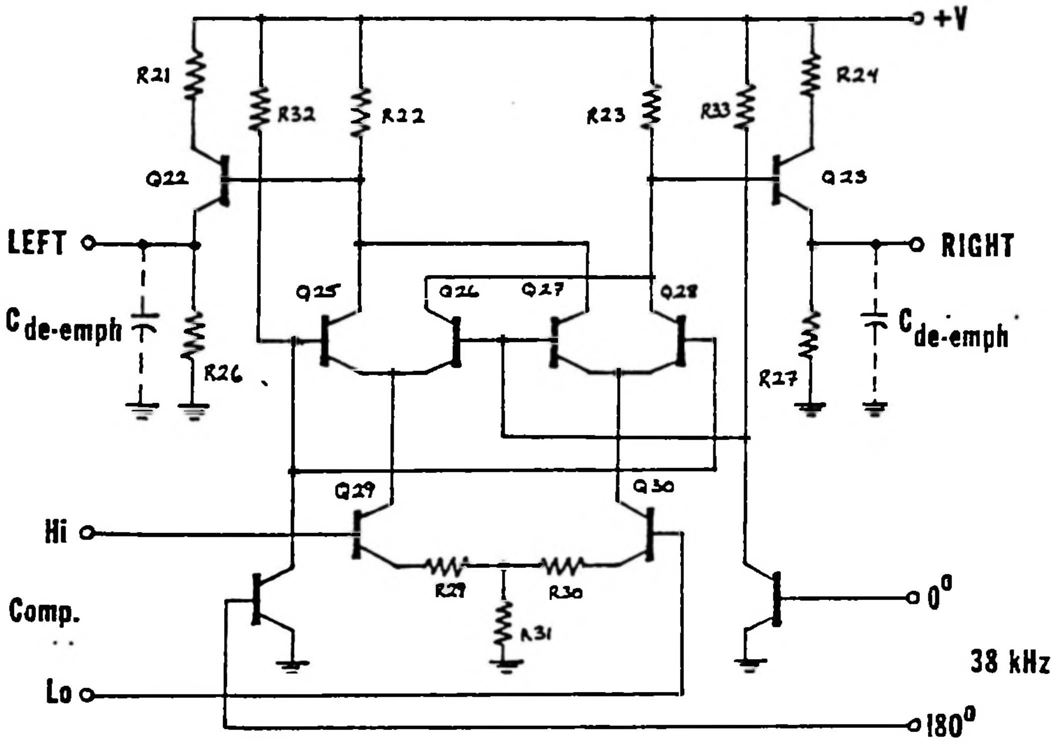
The discrete-component stereo decoder creates noticeable levels of 9 kHz beat-note when 67 kHz SCA is transmitted because the output matrix diodes will also detect signals at twice the switching frequency, or 76 kHz. Thus, 67 kHz is a single sideband 9 kHz off center, producing the beat-note about 6 dB below a double-sideband signal. (It is notable that since this 9 kHz sideband is detected by complementary diode pairs, all signals including the beat-note are out of phase at the Left and Right outputs. This explains why summing the channels completely cancels the beat-note.)

Since detection of the L-R diodes is amplitude-linear, it follows that the level of the 9 kHz beat-note is directly related to the injection, or modulation of the FM carrier by the 67 kHz subcarrier. If one band-pass filters the L-R signal before it is fed to the diodes, or band-stop filters the 67 kHz region of the baseband, there is a reduction in beat-note level commensurate with the relative attenuation of the filter at f_{SCA}.

The sample circuit uses C1 and L4 as a series-resonant shunt arm. This arm traps out part of the baseband around 67 kHz and effectively reduces the 9 kHz beat-note. The sample tuner in this test uses both series and shunt traps to reduce the 67 kHz carrier more than 14 dB. All cost-effective filters of these types compromise stereo separation since they introduce phase shift in the adjacent upper L-R sidebands, particularly at high audio frequencies. It is an acceptable side-effect, considering the apparently large number of older tuners using this technique.

Figure 5

Phase-Locked Loop IC Stereo Decoder, Simplified



The Phase Locked Loop stereo decoder is commonly a monolithic integrated circuit which has been widely used in FM receivers since the mid-1970s. This IC decodes the Left and Right audio channels from the same L+R, Pilot and L-R signals of the stereo FM baseband. Its low number of required external components (usually without inductors), low cost and high audio performance have made it the choice of receiver designers across the consumer price range.

Fortunately for SCA services, PLL stereo decoders have inherently high supression of 67 kHz interference. They were introduced at an especially crucial time, when the use of FM by the public and their expectations of aural quality rose dramatically. It appears to this writer that SCAs would not have enjoyed near the same (albeit still limited) level of acceptance had the earlier discrete system remained in use.

The PLL decoder differs in two basic ways from the discrete decoder: the use of a phase detector, Voltage Controlled Oscillator and divider to regenerate the 38 kHz subcarrier; and a balanced switch which decodes the Left and Right channels without matrixing. The latter technique is primarily responsible for the greatly improved 67 kHz suppression and will be discussed.

The balanced demodulator shown in Figure 5 is typical of the output section of most IC stereo decoders. It is a doubly-balanced multiplier or mixer, and is much the same as the phase comparator in general purpose analog phase locked loops.

Composite baseband signals (including an SCA, if present) are fed to the base of Q29. The same signals, in phase but much lower in amplitude, are fed to the base of Q30 to cancel the crosstalk components between Left and Right outputs that are inherent to the transistor switches Q25 and Q26. Synchronous decoding of the Left and Right channels occurs when the 38 kHz drive signal alternately switches Q25 and Q26 into conduction.

Since the composite baseband signals at the emitters of Q25 and Q26 are fed to the two output stages on complementary half cycles of the 38 kHz subcarrier, even-order multiples of 38 kHz are balanced and produce no audio output. (This explains why PLLs which use this type of phase comparator can only lock to input signals that are odd harmonics of the free-running VCO frequency.)

For a 67 kHz input to the PLL stereo decoder, the demodulated product is:

$$38 \text{ kHz} - 67 \text{ kHz} = 29 \text{ kHz},$$
$$(3\text{x}38 \text{ kHz}) - 67 \text{ kHz} = 47 \text{ kHz, etc.}$$

These frequencies are above audibility and apparently are filtered and de-emphasized sufficiently to avoid side-effects in following systems.

ANALYSIS OF SCA TEST SIGNALS

The results of the SCA subcarrier test frequencies are shown in the next series of figures. Figure 6 shows the output of the original discrete stereo decoder when the 67 kHz subcarrier is frequency modulated to 2.5 kHz peak deviation by a 100 Hz sinewave. What is decoded is the normal bandwidth subcarrier spectrum, centered at 9 kHz. This deviation of the carrier produces a warbling of the 9 kHz beat-note, giving it the term "birdie." Because the energy of the carrier has been distributed among many sideband pairs, the displayed level is lower than at rest. The beat-note product for the PLL decoder is below the -80 dB baseline of the analyzer and is not shown. The tilt of the sideband distribution is caused by the 75 uSec de-emphasis in the stereo decoder.

Figure 6

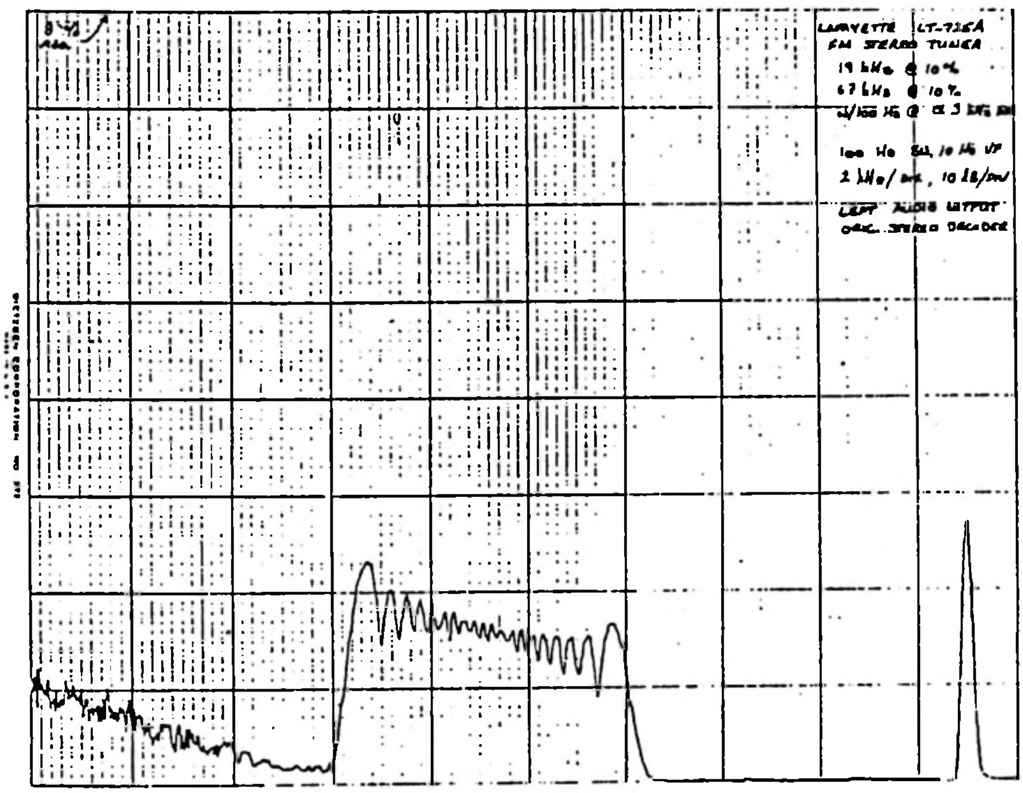


Figure 7 shows the spectrum out of the discrete stereo decoder with a 76 kHz SCA modulated as above. The beat-note product has moved down to a center frequency of 0 Hz. The upper sidebands have been "folded over" the lower ones causing the rough spectral envelope. The average level of the beat-note is about 15 dB higher than the test above, mostly due to the lack of de-emphasis below about 2 kHz. This signal was far more annoying to hear than the 9 kHz beat-note. It was discarded from the field tests because of its audibility, its failure to be removed by 67 kHz traps, and its central position in the upper baseband, which leaves too little bandwidth for another SCA subcarrier to co-exist.

Figure 7

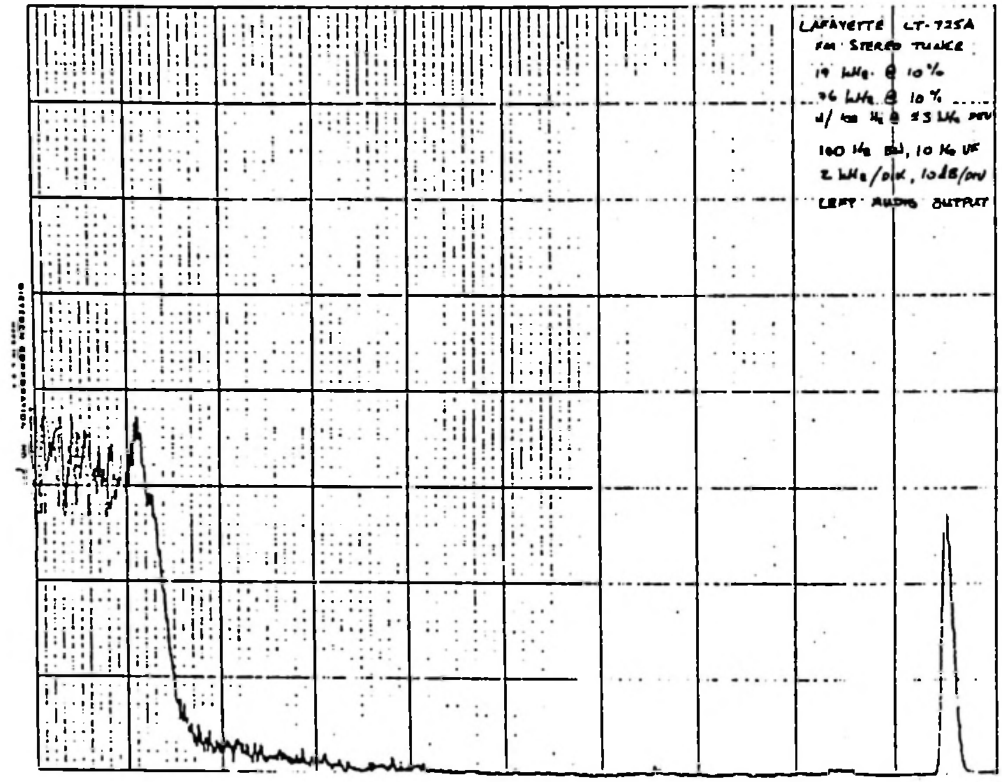


Figure 8

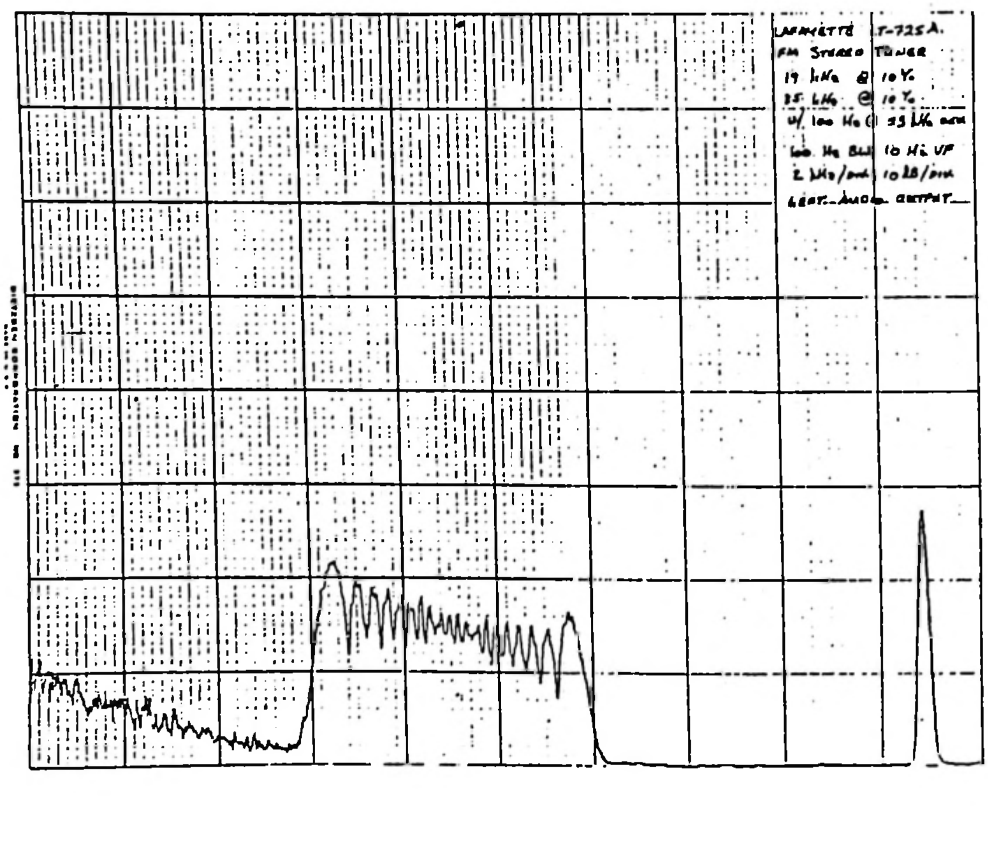


Figure 8 was again produced by the 2.5 kHz modulated SCA subcarrier, moved up to 85 kHz. This frequency is 9 kHz above the crucial second harmonic of 38 kHz, and produces the expected 9 kHz birdie. It is at a level almost exactly that of the 67 kHz subcarrier in the test of this discrete stereo decoder. We can expect this beat-note to be worse than the beat-note for 67 kHz, which would be partially trapped out. It is also insufficiently separated from 67 kHz to permit normal (4 to 6 kHz) deviation of either subcarrier.

Figure 9 shows the result of moving the subcarrier to 95 kHz. This frequency has been proposed in the Quadraphonic Docket. 3/ The subcarrier has moved further from 76 kHz, with the predictable result of moving the beat-note higher still. Since the beat-note is 19 kHz (95 kHz - 76 kHz), it is centered about the 19 kHz Pilot leakage. This signal would presumably be inaudible to listeners, even if it is not filtered out of the audio.

Figure 9

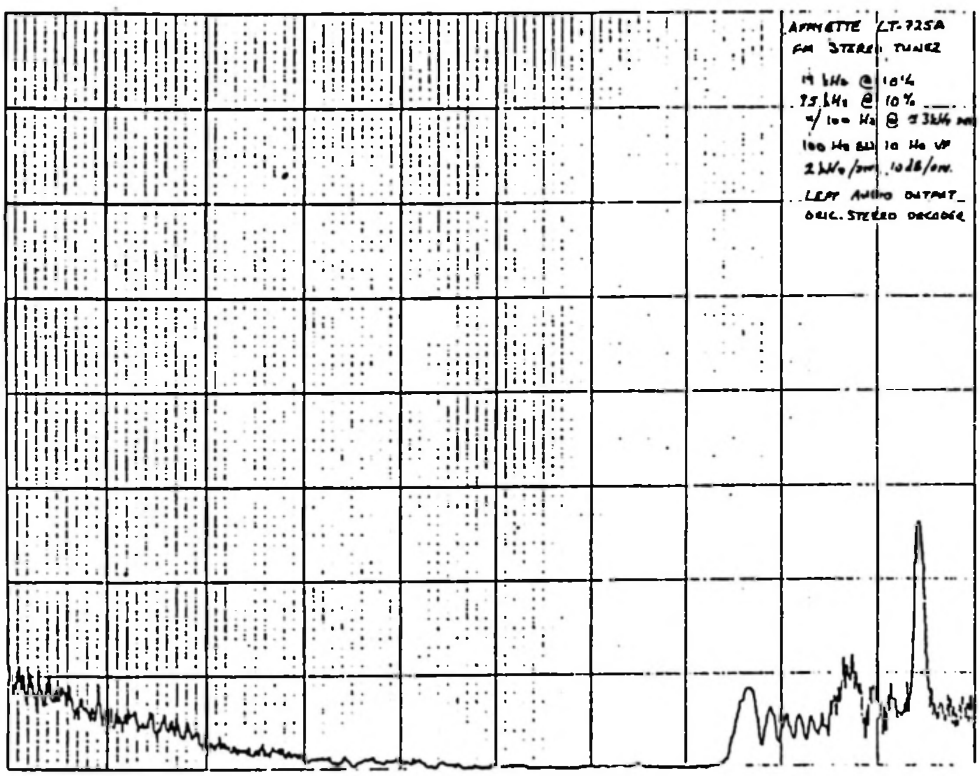


Figure 10

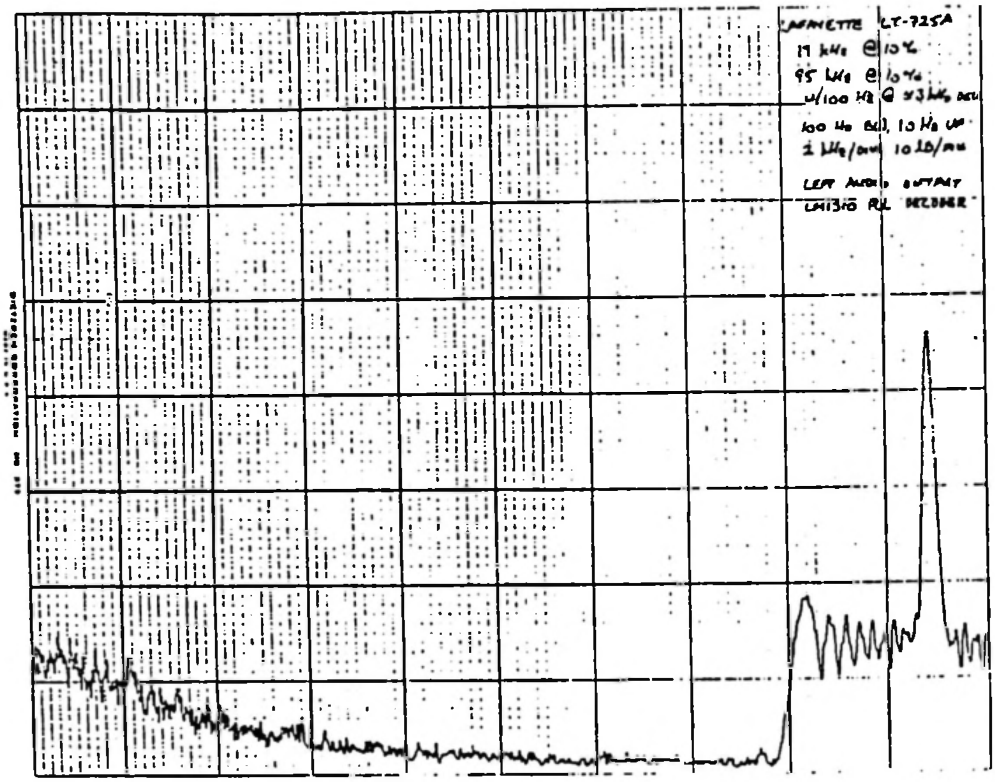


Figure 10 shows the left channel audio spectrum from the PLL stereo decoder. Here again, a modulated SCA carrier is centered about 19 kHz (with a large amount of Pilot leakage). Unlike the discrete decoder, the beat-note results from its frequency relationship with the third harmonic of the 38 kHz switching signal:

$$114 \text{ kHz } [3 \times 38 \text{ kHz}] - 95 \text{ kHz} = 19 \text{ kHz}$$

The amplitude of the beat-note is quite high for this PLL decoder (about -48 dB unmodulated referred to peak L+R level). The beat-note may be removed by the multiplex filters in many consumer receivers and tape recorders. This finding suggests that any spurious signals in an FM station's baseband near 114 kHz may produce a noticeable beat-note for listeners with PLL decoders.

Since the ear can detect coherent signals below noise, such a signal may be audible but not additive on noise meters measuring the full audio bandwidth (15 kHz).

<u>Figure 11</u>

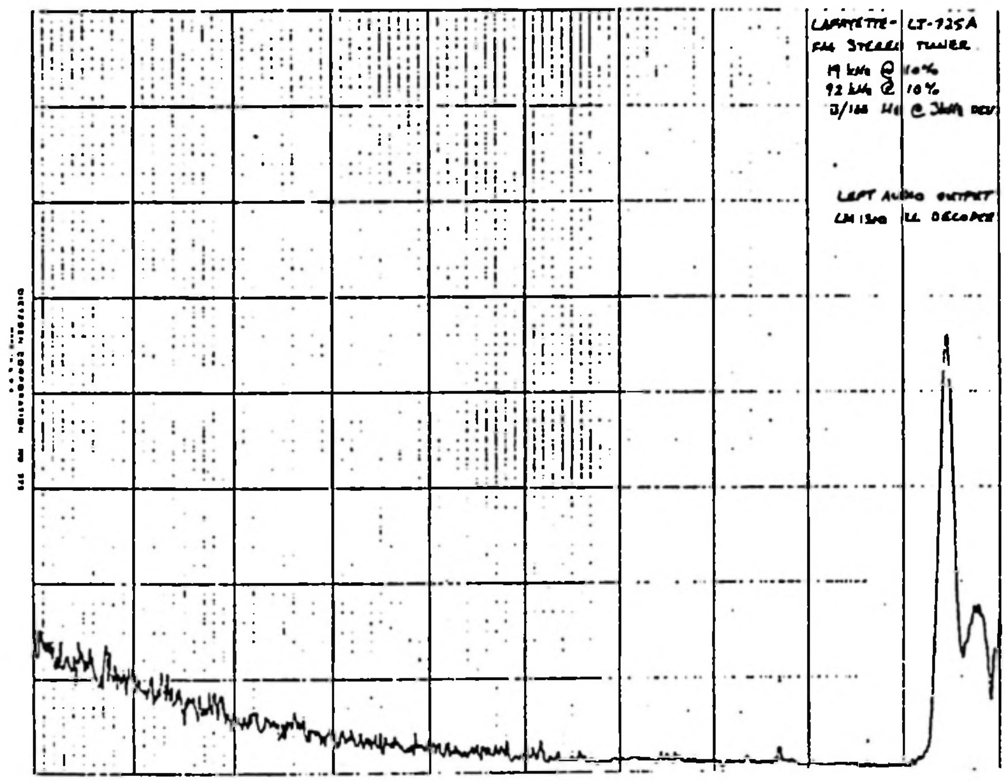


Figure 11 was measured with a 92 kHz subcarrier, showing that the beat-note has almost moved outside the spectrum display. Only the lower edge of the subcarrier is visible.

Figure 12 shows the same 92 kHz subcarrier at the output of the original, discrete stereo decoder. The beat-note is centered at 16 kHz, and in this modulated case has sidebands extending approximately from 13 kHz to 18 kHz. This signal is just perceptible at high listening levels as a very high-pitched buzz. In the opinion of this writer and several others who auditioned the test, this 16 kHz beat-note is much less noticeable than the 9 kHz beat-note resulting from the 67 kHz subcarrier, even when the normal 67 kHz traps were used. This may be due to several reasons:

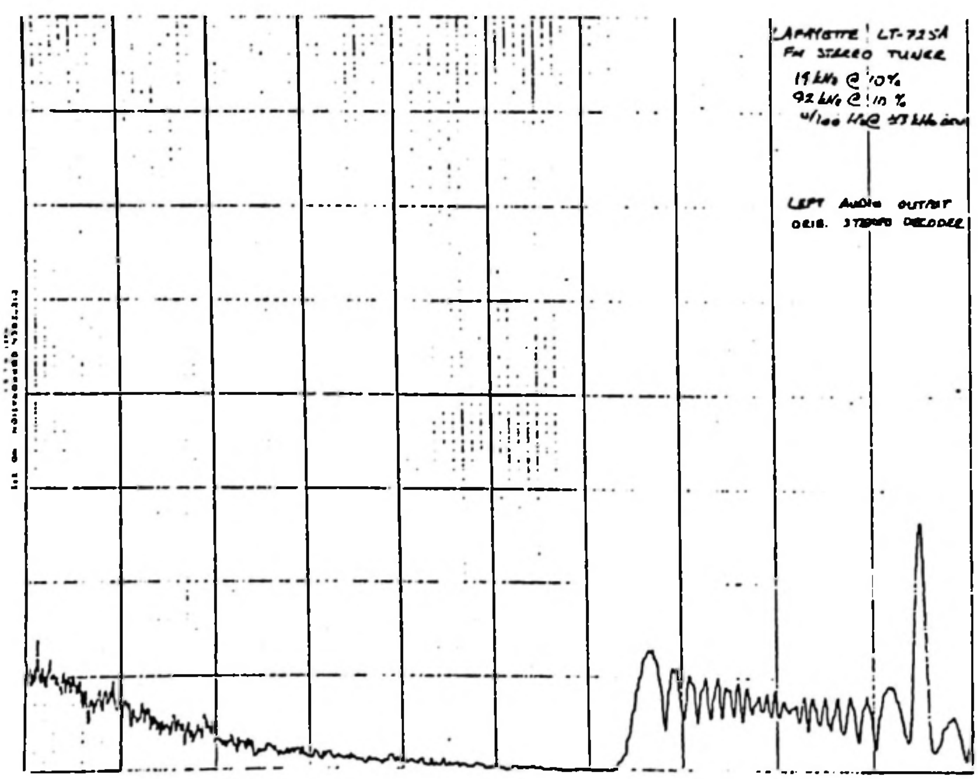
- the ear's sensitivity reduces with increasing frequency, particularly at low listening levels;

- the de-emphasis further reduces the beat-note level; and
- even though instantaneous sidebands from normal program modulation may fall as low as 9 or 10 kHz, the audibility of these occurences was found to be low.

Other advantages of the 92 kHz subcarrier are:

- it is sufficiently removed from 67 kHz to permit both subcarriers to be fully modulated without mutual interference;
- it is lower in the baseband, allowing the subcarrier to be modulated to a higher degree than 95 kHz without instantaneous sidebands exceeding 99 kHz, as suggested in the Commission's Quadraphonic Proposed Rule Making. (±4 kHz maximum deviation for 95 kHz, ±7 kHz for 92 kHz, an increase of 4.9 dB)

Figure 12



OTHER STEREO/SCA BEAT-NOTE CHARACTERISTICS

The previous discussion dealt with interference from the SCA subcarrier generated within the stereo decoder. While this appears to be the prevalent source of interference, another mechanism exists--distortion and intermodulation of signals within the baseband, due to multi-path interference and phase non-linearities in the RF or IF signals.

It has been shown by tests 4/ at NHK (Japan Broadcasting Corporation) that second-order intermodulation of the Pilot and SCA subcarrier occurs when a delayed RF carrier is mixed with the desired RF carrier. This distortion takes the form:

$$f_{SCA} \pm f_{Pilot} = f_{IM\ Product}$$

In the case of 67 kHz or 92 kHz subcarriers, significant products are:

67 + 19 = 86 kHz,
67 - 19 = 48 kHz,
92 + 19 = 111 kHz,
92 - 19 = 73 kHz

While these IM products are not directly audible, some may be once they are demodulated by a discrete stereo decoder:

86 - 76 = 10 kHz (67 kHz subcarrier)
48 - 38 = 10 kHz (67 kHz subcarrier)
111 - 76 = 3 kHz (92 kHz subcarrier)
73 - 76 = 3 kHz (92 kHz subcarrier)

and a PLL stereo decoder:

48 - 38 = 10 kHz (67 kHz subcarrier)
111 - 114 = 3 kHz (92 kHz subcarrier)

In addition to the above products, second order IM could result from the two SCA subcarriers.

92 - 67 = 25 kHz;

which would demodulate in either type of stereo decoder as:

25 - 38 = 13 kHz

Figure 13 shows the transmission equipment at WETA-FM, as used for the field tests and the next four graphs.

Figure 13

WETA-FM Transmission System

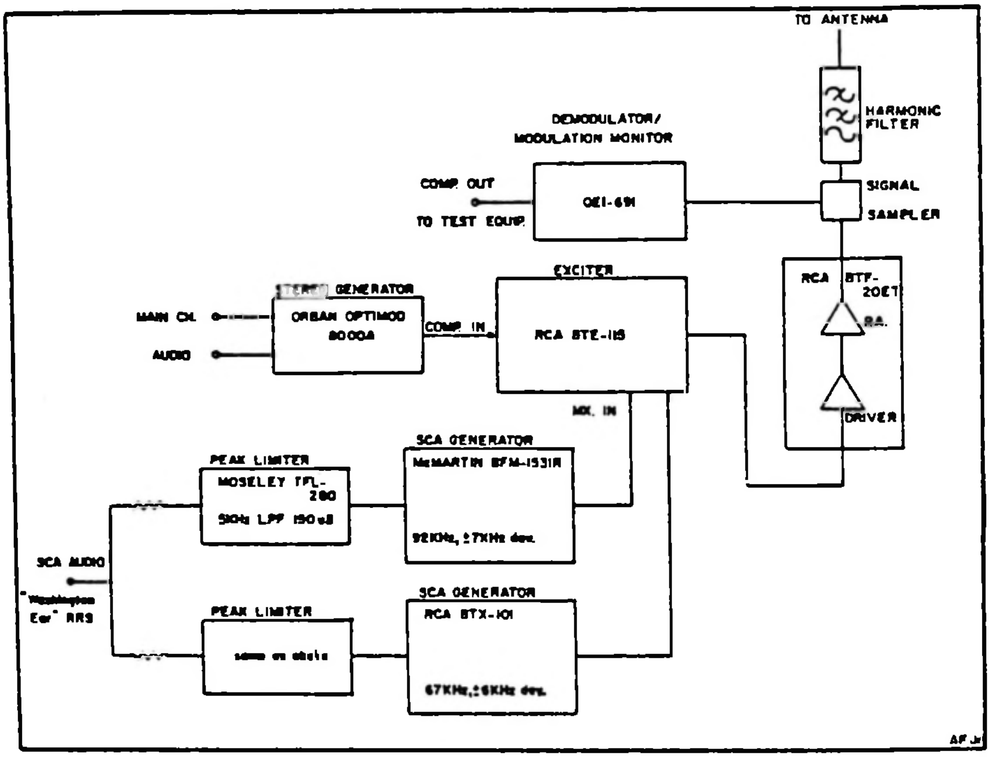
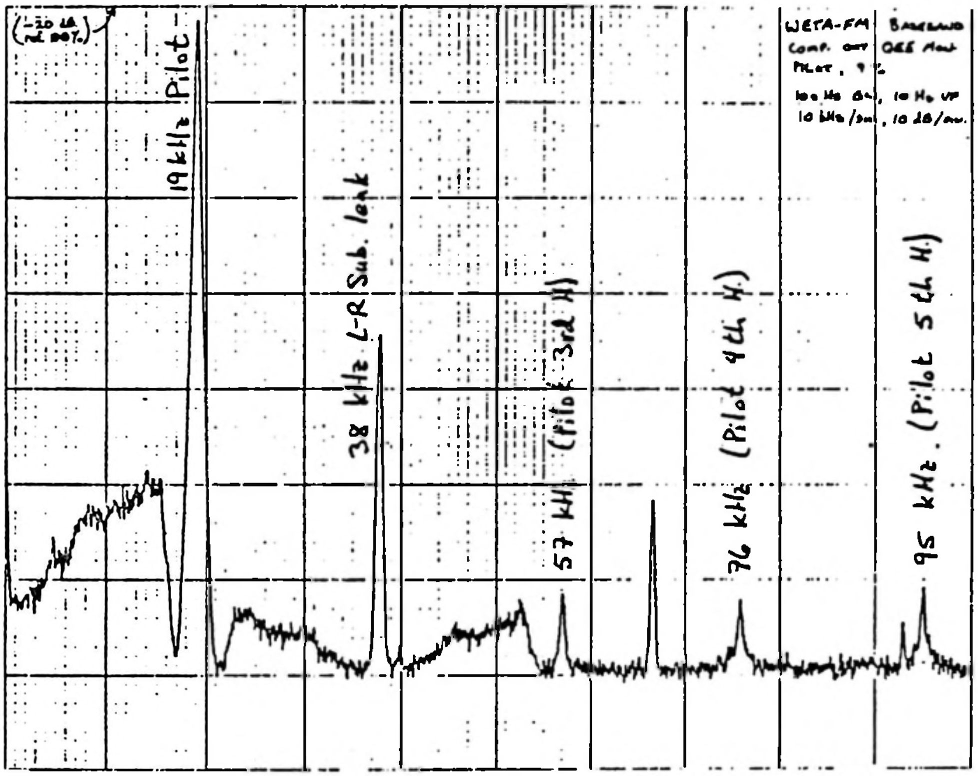


Figure 14 shows the demodulated composite baseband of WETA-FM with no subcarriers (67 kHz is leaking in at -72 dB referred to 75 kHz deviation). The Pilot and supressed 38 kHz carrier are visible. Figure 15 shows the same spectrum with the addition of spurious signals at 48 kHz and 86 kHz. A very low level signal at about 81 kHz appeared but remains identified. Figure 16 shows the spectrum with both 67 kHz and 92 kHz subcarriers. Two more spurious signals have been introduced, at 25 kHz (-78 dB) and 73 kHz (-69 dB) and 111 kHz (not shown, and also -69 dB).

Figure 14



The level of most beat-note signals after stereo demodulation may be predicted by the following equation:

$$L_{beat} = L_{spur} - D_{de\text{-}emph.} - 6 \text{ dB}$$

Where L_{spur} is the level of the spurious IM product, $D_{de\text{-}emph.}$ is the de-emphasis value in dB for the resultant beat-note frequency and 6 dB corrects for the existence of only one sideband.

Using the 95 kHz subcarrier shown in Figure 10, we can verify the beat-note level for 114-95 kHz:

$$L_{beat} = -20 \text{ (10\% injection)} - 19 \text{ (for 19 kHz)} - 6$$
$$= -45 \text{ dB}.$$

Adding 18 dB to correct for the difference between the displayed beat-note and a unmodulated carrier (as shown in Figures 3 and 5) and 3 dB for the amplitude rolloff of the IF detector at 95 kHz:

$$-45 - 18 - 3 = -66 \text{ dB}$$

which agrees with the measured level shown in Figure 10.

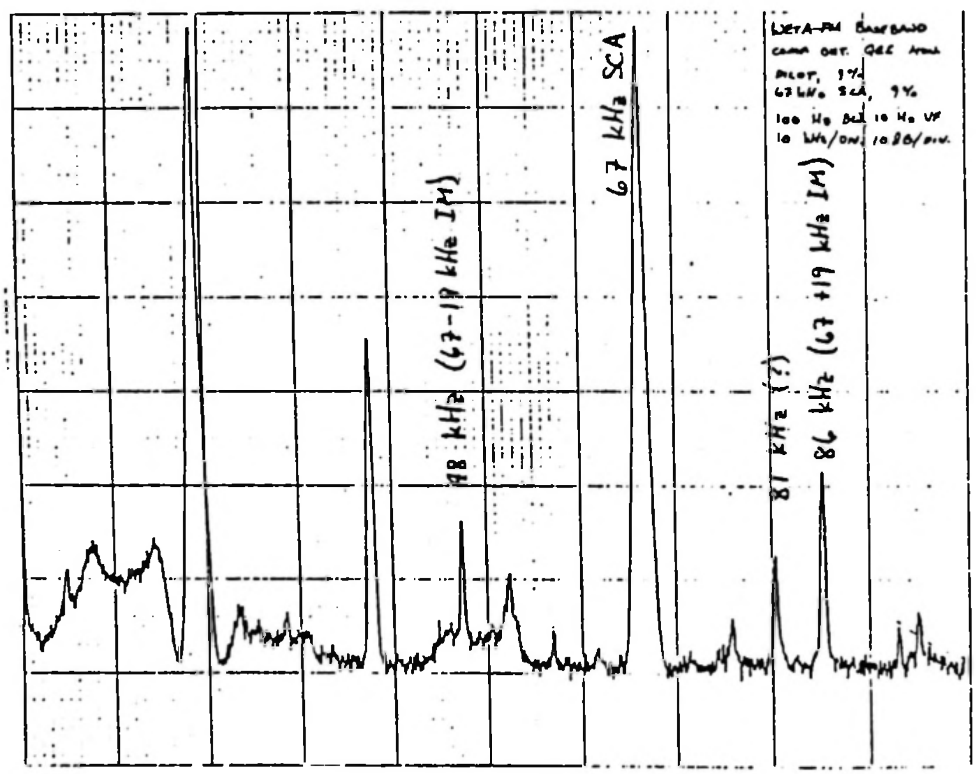
Estimating the level of a 3 kHz beat-note that would result from the high-side mix of the 92 kHz subcarrier and the Pilot (92 + 19 = 111 kHz) in the transmission system, we have:

$$\begin{aligned} L_{beat} &= -69 - 3 \text{ (for 3 kHz)} - 6 \\ &= -78 \text{ dB.} \end{aligned}$$

Estimates for discrete component stereo decoders for spurious signals near 76 kHz are about 10 dB lower than those given in the above equation.

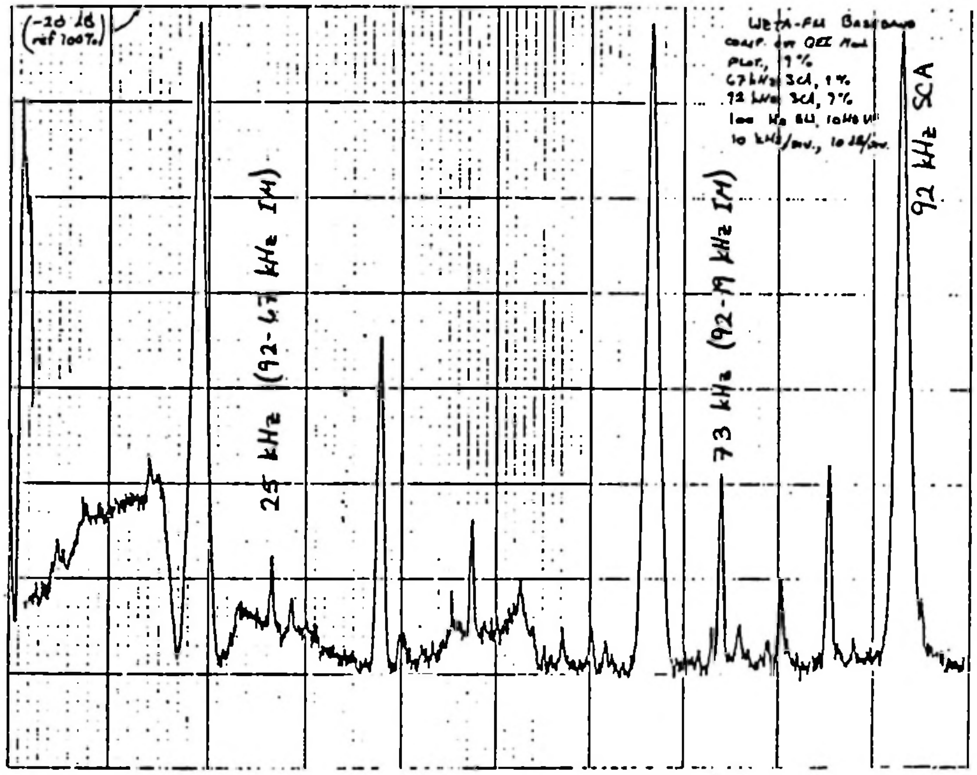
Because of noise masking, stereo listeners were usually unable to discern beat-notes of IM products resulting from multipath distortion at levels below -65 dB.

Figure 15



Hotizontal: 0-100 kHz, 10 kHz/Div., Vertical: 10 dB/Div., top line is -20 dB referred to 100% modulation.

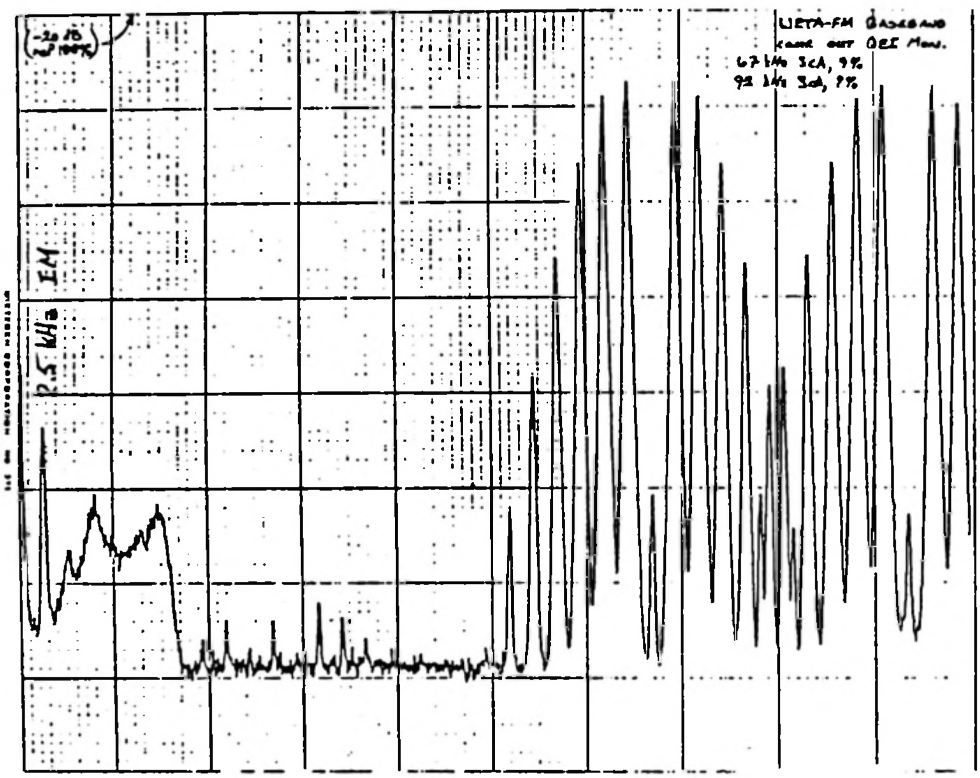
Figure 16



On the basis of careful off-air listening tests over a period of two months, the above two sets of possible demodulated signals were evident only in the presence of significant amounts of multipath. It was found that the mix between two unmodulated SCA subcarriers was more prevalent than between either carrier and the stereo Pilot. In almost all cases, however, the resulting beat-note was below the stereo noise floor and not easily measurable. Elaborate tests would be needed to document beat-note levels for various desired/undesired bi-path levels and delay times. Since no listener complaints were noted during the entire test, including extended programs of low-level classical music, it can be assumed that the possible interference resulting from an additional SCA at 92 kHz was negligible.

A final comment on the possible SCA interference mechanism: sideband energy of the two subcarriers should not overlap on a frequent basis. Figure 17 shows the result of modulating both SCAs with 2495 Hz at 6 kHz. A direct mix occurs between the upper sidebands of 67 kHz and lower sidebands of 92 kHz, causing a 2495 Hz IM product. This signal was quite audible in a mono receiver. These conditions are fortunately non-existent when at least one subcarrier is modulated by program audio, since the energy is held far closer to the carrier and the duty-cycle of sideband overlap is extremely low.

Figure 17



92 kHz SCA FIELD TESTS

Subjective and instrument measurements of the 92 kHz subcarrier were made with an FM/SCA receiver system shown in Figure 18. The system is actually two standard table-model SCA radios of the type used by individual listeners. To make received performance for the two subcarriers comparable, a transistor emitter follower was added to one radio to drive both subcarrier detectors from the same composite baseband.

Figure 18

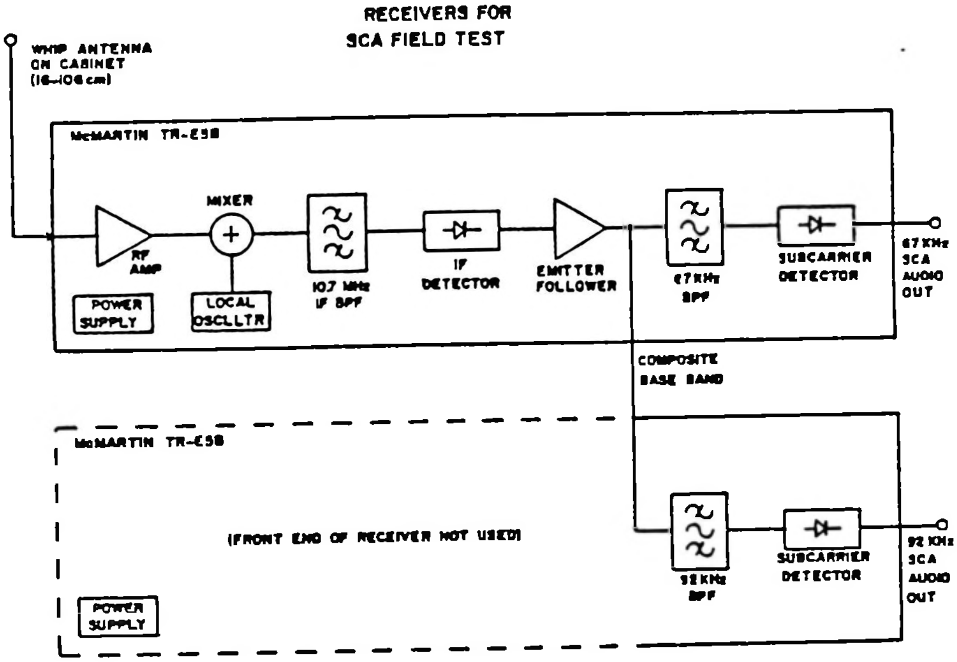
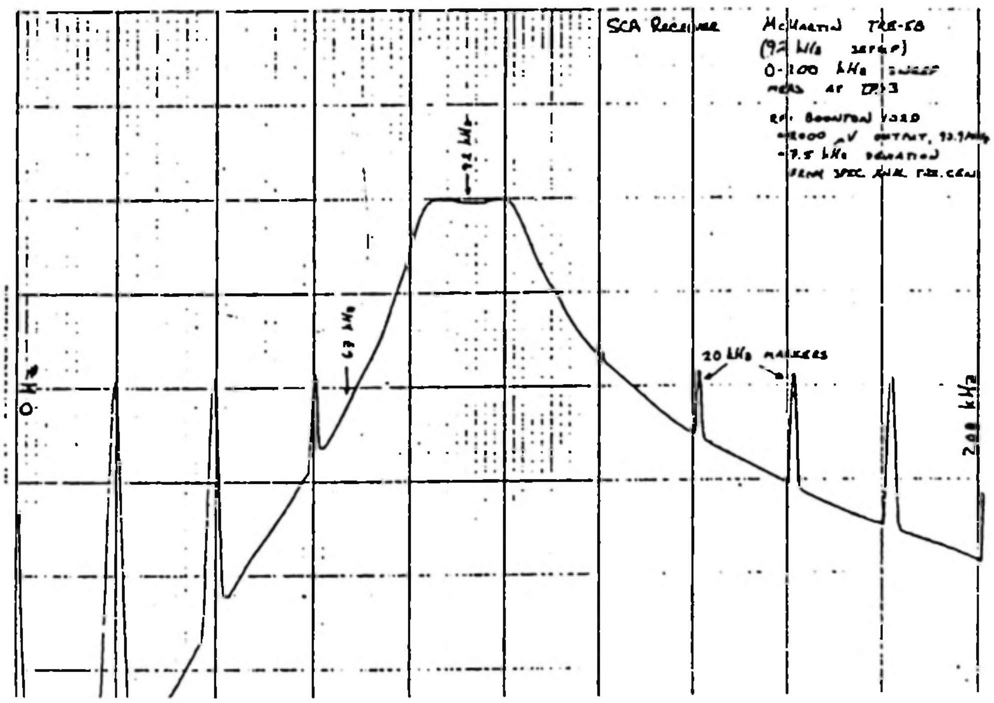


Figure 19 shows swept response of the 92 kHz bandpass filter, which was adjusted to produce a 3 dB bandwidth of 12 kHz. The 67 kHz filter had a similar shape. Overall frequency response was checked

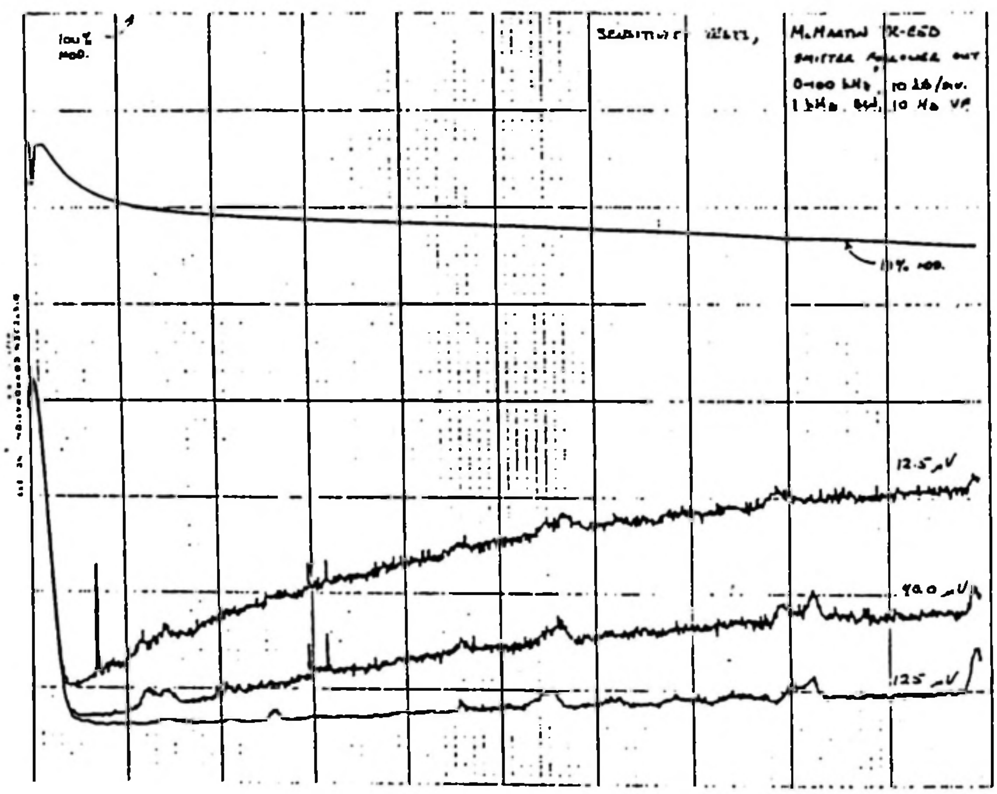
Figure 19



for the subcarrier detectors. Both had a response within ±1 dB of the 150 uSec de-emphasis curve to 3 kHz and was down 2 dB from the nominal curve at 5 kHz, measured at the input to the associated SCA generator.

Figure 20 shows the noise level for three levels of RF delivered to the SCA receiver antenna terminals (50 ohms impedance), and the swept response of the IF detector, both taken at the output of the emitter follower.

Figure 20



It is notable that the noise level does not rise at a constant 6 dB per octave rate [5/] through the entire baseband, nor does it increase 10 dB for each 10 dB reduction in RF level. The former is due to the amplitude response limitations of the IF filter and detector. The latter is limited by internal noise of the receiver and test generator.

To determine theoretical Signal-to-Noise ratios for the subcarriers, we must first calculate the carrier to noise ratio from the spectrum analyzer's display: [6/]

$$C/N = C/N_{S.A.} + CF - 10 \log_{10} \left(\frac{B_{SCA}}{B_{S.A.}}\right)$$

$$C/N_{S.A.} + 2.5 - 10 \log_{10} \left(\frac{12}{1}\right)$$

$$C/N_{S.A.} + 2.5 - 10.8 = C/N_{S.A.} - 8.3 \text{ dB}$$

where C/N is the carrier to noise ratio in dB, B_{SCA} is the predetection bandwidth of the SCA receiver in kHz, $B_{S.A.}$ is the spectrum analyzer filter bandwidth in kHz, CF is the correction factor for the way the analyzer displays noise in the logarithmic mode and $C/N_{S.A.}$ is the analyzer's displayed ratio of carrier to averaged noise in dB.

Using the 10% modulation vs. frequency sweep line, we have a $C/N_{S.A.}$ at 12.5 uV of:

51.5 - 22.5 = 29 dB for 67 kHz; and
49.5 - 23.5 = 26 dB for 92 kHz

The carrier to noise ratios are:

29 - 8.3 = 20.7 dB for 67 kHz; and
26 - 8.3 = 17.7 dB for 92 kHz
(a difference of -3 dB)

This is very close to the theoretical noise rise, given by:

$$20 \log \left(\frac{92 \text{ kHz}}{67 \text{ kHz}}\right) = 2.75 \text{ dB}.$$

The FM improvement for the frequency-modulated SCA is:

$$FM = 10 \log_{10} \left(1.76 \times \left(\frac{B_{SCA}}{B_{AF}}\right) \times \left(\frac{f}{B_{AF}}\right)^2\right) + I_d$$

where B_{AF} is the SCA audio bandwidth, f is the peak deviation and I_d is the de-emphasis improvement. Choosing both SCA subcarriers to have 5 kHz audio bandwidth, a nominal deviation of 6 kHz, and an I_d of 8.8 dB (resulting from 150 uSec de-emphasis), the FM improvement is:

$$FMI = 10 \log_{10} \; 1.76 \times \frac{12}{5} \times \frac{6}{5}^2 + 8.8 = 7.8 + 8.8 = 16.6 \text{ dB}$$

The predicted SCA signal-to-noise ratio at the displayed noise level of 12.5 uV is:

$$S/N = C/N + FMI$$
$$= 20.7 + 16.6 = 37.3 \text{ dB for 67 kHz; and}$$
$$= 17.7 + 16.6 = 34.3 \text{ dB for 92 kHz}$$

The actual measured signal to noise ratios were:

31.5 dB for 67 kHz; and
30.0 dB for 92 kHz

The actual S/N was poorer than predicted by:

-5.8 dB for 67 kHz; and
-4.3 dB for 92 kHz

This difference is due mainly to dissimilar filter shapes in the SCA receivers and spectrum analyzer and non-ideal detector characteristics.

The measured signal-to-noise ratio of the 92 kHz SCA here is 1.5 dB less than 67 kHz. However, the difference can be minimized when a peak deviation of 7 kHz is used, an increase of 1.3 dB over the 6 kHz deviation used at 67 kHz. (For music transmission, a slightly greater pre-detection filter bandwidth is suggested if lower levels of audio IM distortion are desired.)

A tape recording was made during the field tests comparing the two subcarriers while fed by identical programming, using only the receiver's telescoping whip antenna. A variety of locations were chosen to include strong, moderate and weak signal conditions, multi-path reception and an RF cross-modulation from stronger carriers. Aside from more noticeable impulse noise (automobile ignition radiation, for example), subjective performance at all sites was almost identical. Impulse noise may be higher since the peak energy of the noise should be 3 dB higher at 92 kHz, and if it exceeds the subcarrier level will momentarily overwhelm the detector. This situation occured mostly with a weak RF input and apparently strong sources of noise (which were even slightly audible from the main channel detector).

In bench texts, the maximum crosstalk into the sub-channel was -56 dB at 92 kHz and -54 dB at 67kHz with 90% modulation of the main channel. The principal crosstalk component was a second harmonic of the modulating sinewave. The full-quieting signal-to-noise ration of both receivers was approximately 53 dB, referred to 6 kHz modulation.

CONCLUSION

Several conclusions may be drawn from the test results which may be useful to stations operating or considering new SCA services:

-- The principal source of SCA/Stereo interference in older FM receivers is caused by the beat-note between two times the 38 kHz subcarrier and the instantaneous SCA frequency, and occurs only within the stereo decoder;

-- Modern (less than 10 years old) FM receivers inherently have low amounts of SCA/Stereo interference at all subcarrier frequencies between 53 and 99 kHz, do to the use of Phase Locked Loop stereo decoders and improved IM distortion performance;

-- Non-linearities in the RF domain (such as inadequate exciter, transmitter and antenna bandwidth or symmetry, standing waves in transmission lines, multipath reception, receiver misalignment and user mistuning) contribute to second order IM distortion of the baseband signals, the products of which may cause small amounts of audible beat-notes from the stereo decoder; main-to-SCA crosstalk will also increase;

-- FM modulation monitors approved for station use with a baseband of 75 kHz may be inaccurate for measuring higher subcarriers (amplitude rolloffs of up to 10% at 100 kHz were noticed on a sampling of current units);

-- A new SCA subcarrier should be placed between 92 and 95 kHz because it will not produce noticeable beat-notes in either old or new stereo decoders;

-- 92 kHz is preferred as a new subcarrier frequency because it allows a higher modulation index (peak deviation/audio bandwidth) than possible at 95 kHz, which makes overall performance similar to that at 67 kHz;

-- There is sufficient bandwidth between 53 and 99 kHz to permit the simultaneous operation of two SCA subcarriers;

-- Operation of SCA subcarriers at 67 and 95 kHz created no complaints or comments of interference from WETA-FM staff or listeners during the two-month duration of the tests, despite many music programs of wide dynamic range on the main channel.

ACKNOWLEDGEMENT

The author wishes to thank the Greater Washington Educational Telecommunications Association, Inc. for securing the Experimental Authorization and providing the transmission facilities for the test sub-carriers; McMartin Industries, Inc. of Omaha, Nebraska, for the loan of SCA generators and receivers used in the field tests; Wayne Hetrich, Satellite System Engineering Manager of NPR, for his valuable advice in the preparation of the report; and Ken Sleeman, Maintenance Engineer of WETA-FM for setting up and operating the transmission equipment and cheerfully helping on tests at unreasonable hours of the morning.

REFERENCES

1/ Federal Communications Commission, FM Quadraphonic Broadcasting Docket, No. 21310, released August 14, 1980, pg. 22.

2/ Discussions of SCA Receiver Standards Committee (funded by the Corporation for Public Broadcasting), held in 1975, not published.

3/ Ibid., #1.

4/ Mitsuo Ohara, NHK Technical Research Laboratories, "Distortion and Crosstalk caused by Multipath Propogation In Frequency-Modulation Sound Broadcasting," IEE Transactions On Broadcasting, Vol. BC-26,. No. 3, September, 1980, pp. 79-80.

5/ Murray G. Crosby, "Frequency Modulation Noise Characteristics," Proc. I.R.E., Vol. 25, No. 4, April, 1937, pg. 482.

6/ Morris Engelson, Tektronix, Inc., "Noise Measurements Using the Spectrum Analyzer," Application Note AX-3260, 1977, pg. 5.

ISSUES CONCERNING MODULATION LEVELS

DURING FM SCA OPERATION

HARRISON J. KLEIN

WESTINGHOUSE BROADCASTING AND CABLE, INC.

NEW YORK, NEW YORK

Introduction

The FM Subsidiary Communications Authorization (SCA) is one of the most underutilized parts of the FM broadcast spectrum. 1982 FCC data indicate that less than 27% of FM stations use their SCA, and, for all FM stations, the amount of SCA non-use comes to approximately 84,500 hours per day (1).

Yet this lack of SCA utilization does not result from a lack of demand for SCAs, at least in the larger markets. It results primarily from a decision by many FM stations not to make their SCAs available for use, because of a fear that SCA operation would adversely impact main channel programming by degrading the quality of the main channel. Historically, these degradations have included:

1. Reduction in program modulation of approximately 1 dB (10%);
2. Crosstalk of the SCA into the stereo subchannel; and
3. The appearance of beat notes in the received audio.

The latter two problems are no longer significant. The development and widespread use of FM stereo detectors using phase locked loop circuitry has virtually eliminated the generation of audible beat notes by receivers. Beat notes or crosstalk generated by the FM station itself can be eliminated through the use of properly designed and adjusted stereo generators, SCA generators, FM exciters, transmitters, and antennas. Although multipath interference does somewhat increase the amount of SCA-to-stereo and beat note interference, this increase occurs in the presence of an already degraded stereo signal due to multipath. The SCA effects are generally not the primary problem.

However, the modulation issue remains significant. Since the FCC continues to limit total carrier modulation to 100% (75 kHz deviation), addition of an SCA with 10% injection requires a main channel modulation backoff of 10%, or slightly less than 1 dB, in order to prevent overmodulation. While a 1 dB loudness drop is certainly not serious (it is below the threshold of audibility for most people), some broadcasters feel it is significant enough to dissuade them from SCA use. Therefore, any reduction in the amount of main channel backoff required during SCA use would help to eliminate this last remaining disincentive to SCA implementation, and should help to significantly increase SCA utilization.

Occupied Bandwidth Considerations

The main channel modulation backoff is intended to prevent any increase in occupied bandwidth when the SCA is added. But because of the complex nature of the FM modulation process, and because the SCA subcarrier is separated in frequency from the other modulating components, the contribution of a 10% SCA subcarrier to occupied bandwidth does not necessarily equal the contribution of 10% modulation in the main channel. The tests described in this paper were designed to measure the extent to which a reduction in main channel modulation is actually necessary when one or more SCA subcarriers are utilized. The engineering report describing the test results was submitted to the FCC by Westinghouse Broadcasting and Cable, Inc. (Group W) as comments in the Notice of Proposed Rulemaking concerning use of the SCA, Docket No. 82-536.

The test results indicate that the backoff requirements during stereophonic main channel operation should be as follows:

1. When the first SCA or SCAs of up to 10% total injection are utilized, the main channel backoff should be one-half the total injection of the subcarriers. For example, use of a single SCA subcarrier with 10% injection would require a main channel backoff of 5%, for a total modulation of 105%.

2. When additional SCAs are added for a total SCA injection of between 10% and 20%, no additional main channel backoff should be required. For example, use of two SCA subcarriers, each with 10% injection, would still require a main channel backoff of 5%, for a total modulation of 115%.

3. Modulation of greater than 100% should be permitted only when SCAs are in use. Stereophonic and monophonic transmissions without SCAs should continue to be limited to 100% modulation.

It is shown below that the increase in modulation recommended above, to as much as 115%, does not cause any significant increase in occupied bandwidth, nor does it significantly affect the audio performance of FM receivers.

Figure 1 contains a block diagram of the measurement equipment configuration. The performance of the FM exciter, stereo generator, and SCA generators exceeded all requirements of FCC rules.

The First SCA Can Be Added With Only a 5% Main Channel Modulation Backoff

Figure 2a shows an overlay of the RF spectra for stereo-only modulation (no SCA) of 100% versus stereo plus 67 kHz SCA modulation of 100%. The lower trace in figure 2a is for the stereo plus SCA case, showing a decrease in sideband energy when a 9% SCA is used with a full 9% main channel modulation backoff. (National Public Radio used the same overlay method in its comments in Docket No. 82-536, and also noted this sideband energy reduction.)

Figure 2b shows an alternate view of the same two RF spectra: it is the arithmetic difference between the two spectra of figure 2a, where no difference would be indicated by a horizontal line across the center of the screen. It is clear that a full 9% main channel modulation backoff when a 9% SCA is added reduces main channel sideband energy more than necessary to compensate for the sideband energy created by the SCA.

Figure 3a shows an overlay of the same 100% stereo modulation as in figure 2a, versus a stereo plus 67 kHz SCA modulation of 105% total. Figure 3b shows that there is almost no difference between the two spectra, demonstrating that a 5% main channel modulation backoff is appropriate when the first SCA is utilized.

A 67 kHz SCA subcarrier frequency was used in the above measurements, but the results are also valid for higher SCA frequencies. Figure 4a is an overlay of an 82% stereo plus 92 kHz SCA spectrum (100% total modulation), versus an 87% stereo plus 92 kHz SCA spectrum (105% total modulation). Figure 4b shows that there is almost no difference between the two spectra, again demonstrating that 5% is an appropriate main channel backoff for the first SCA.

While the 67 kHz SCA case above did show a change in spectrum shape when going from 100% to 105% modulation, the same modulation increase when using a 92 kHz SCA results in almost no change in spectrum shape. This apparent "something for nothing" result can be explained because the additional sideband energy created by the 5% main channel modulation increase is small compared with the existing sideband energy due to the 92 kHz SCA, and therefore does not significantly change the total.

The Second SCA Can Be Added With No Additional Main Channel Modulation Backoff

Figure 5a shows an overlay of the RF spectra for stereo plus 92 kHz SCA modulation of 100%, versus the same signal with the addition of a 9% injection 67 kHz SCA (no main channel backoff). Figure 5b shows that the addition of the second SCA resulted in no significant increase in sideband energy.

Figures 6a and 6b show a similar condition to figures 5a and 5b, except that the main channel modulation has been increased by 5% for a total modulation with one SCA of 105%. Again, figure 6b shows no significant increase in sideband energy due to the second SCA, demonstrating that no main channel modulation reduction is required when adding the second SCA.

Figure 7 summarizes the effects on occupied bandwidth of the main channel modulation backoff requirements proposed in this paper. Figure 7a is an overlay of a simple stereo plus one SCA, 100% modulation spectrum, versus a stereo plus two SCAs, 114% modulation spectrum. Figure 7b shows only a small increase in sideband energy. Yet this insignificant occupied bandwidth change provides a significant improvement in spectrum efficiency. Not only is a second SCA service added with no penalty to the station's main channel, but the most important inhibiting factor to first SCA use, the 10% main channel backoff, has been substantially lessened.

Although tone modulation was used in figures 2 through 7 to simplify the measurement procedure, the results are also valid for program modulation. Figure 8 is identical to figure 7 except that stereo program modulation was used instead of tone modulation. The program material consisted of "Monday Morning," a popular recording by Fleetwood Mac. The spectrum analyzer was left in the

"MAX HOLD" mode for approximately three minutes. Accumulating spectral peaks for longer than three minutes did not result in any visible change in the displayed spectrum.

The difference display of figure 8b confirms the conclusions reached from figure 7b: even under program modulation conditions, the addition of a second SCA and the increase in main channel modulation by 5% results in no significant increase in sideband energy. As was mentioned above, this "something for nothing" result occurs because the sideband energy created by the main channel modulation increase and second SCA is substantially less than the sideband energy already present from the main channel and first SCA, so its addition changes the total energy very little.

Although the above measurements were made with unmodulated SCA subcarriers, similar measurements were made with the SCAs fully frequency modulated by pink noise. There was no observable difference in the resulting RF spectra from the unmodulated SCA case.

Modulation of Greater Than 100% Should Be Permitted Only When SCAs Are In Use

A maximum modulation limit of 115% under SCA modulation conditions is technically viable without objectionable increases in occupied bandwidth only because of the unique spectral characteristics of low deviation subcarriers. Stereophonic or monophonic FM transmissions without SCAs should continue to be limited to 100% modulation. Figure 9a shows an overlay of a stereo plus two SCAs, 114% modulation spectrum, versus a stereo-only 114% modulation spectrum. Figure 9b shows that stereo-only 114% modulation contains significantly more sideband energy than stereo plus two SCAs modulation of the same amount.

Modulation of Up to 115% Does Not Observably Degrade Receiver Audio Performance

National Public Radio commented in Docket No. 82-536 (2) that operation with a peak deviation of 109% does not result in any significant audio performance deterioration. Distortion remained below residual levels, system noise (stereo or SCA) was generally unaffected, and intermodulation products were very low and not affected by the increase in peak deviation.

A comprehensive series of tests made using a consumer-type Sony ST-A30 FM stereo tuner confirms these results. Measurements were made of frequency response, harmonic and intermodulation distortion, signal-to-noise ratio, and stereo separation, under the eleven different conditions shown in table 1. There was no observable deterioration in performance as the peak deviation increased. Although other receivers could conceivably yield different results, there is no reason to expect that performance of existing receivers will be significantly degraded by operating with two SCAs and 115% total modulation.

Conclusion

The measurements described above indicate that the modulation limit of an FM station during operation with one 10% SCA can be increased to 105% with no significant change in occupied bandwidth or receiver performance. If two 10% SCAs are used, the modulation limit can be increased to 115%. Under these conditions, an FM station must only reduce stereo program level by less than 0.5 dB when using up to two SCAs. This amount is truly insignificant by any reasonable standard, and would thus eliminate the one remaining technical obstacle to wider use of SCAs by FM stations. It is hoped that the FCC will implement these revised modulation limits when Docket No. 82-536 is acted upon.

Acknowledgments

Grateful thanks is extended to Broadcast Electronics Inc., Quincy, Illinois, for providing the equipment and facilities used to make these measurements.

References

1. Notice of Proposed Rule Making in BC Docket No. 82-536, FCC 82-368, 47 Fed. Reg. 36235 (August 19, 1982), paragraphs 8, 14.

2. Comments of National Public Radio in BC Docket No. 82-536, page 30, 32.

TABLE 1

Receiver Performance Measurement Conditions

(Note: U = Unmodulated, M = Modulated with pink noise)

Modulation Percentage

Baseband Component	Test 1	Test 2	Test 3	Test 4
Stereo	91	82	82	82
Pilot	9	9	9	9
67 kHz SCA		9U	9M	
92 kHz SCA				9U
Total	100	100	100	100

	Test 5	Test 6	Test 7	Test 8
Stereo	82	87	87	87
Pilot	9	9	9	9
67 kHz SCA		9U	9M	
92 kHz SCA	9M			9U
Total	100	105	105	105

	Test 9	Test 10	Test 11
Stereo	87	87	87
Pilot	9	9	9
67 kHz SCA		9U	9M
92 kHz SCA	9M	9U	9M
Total	105	114	114

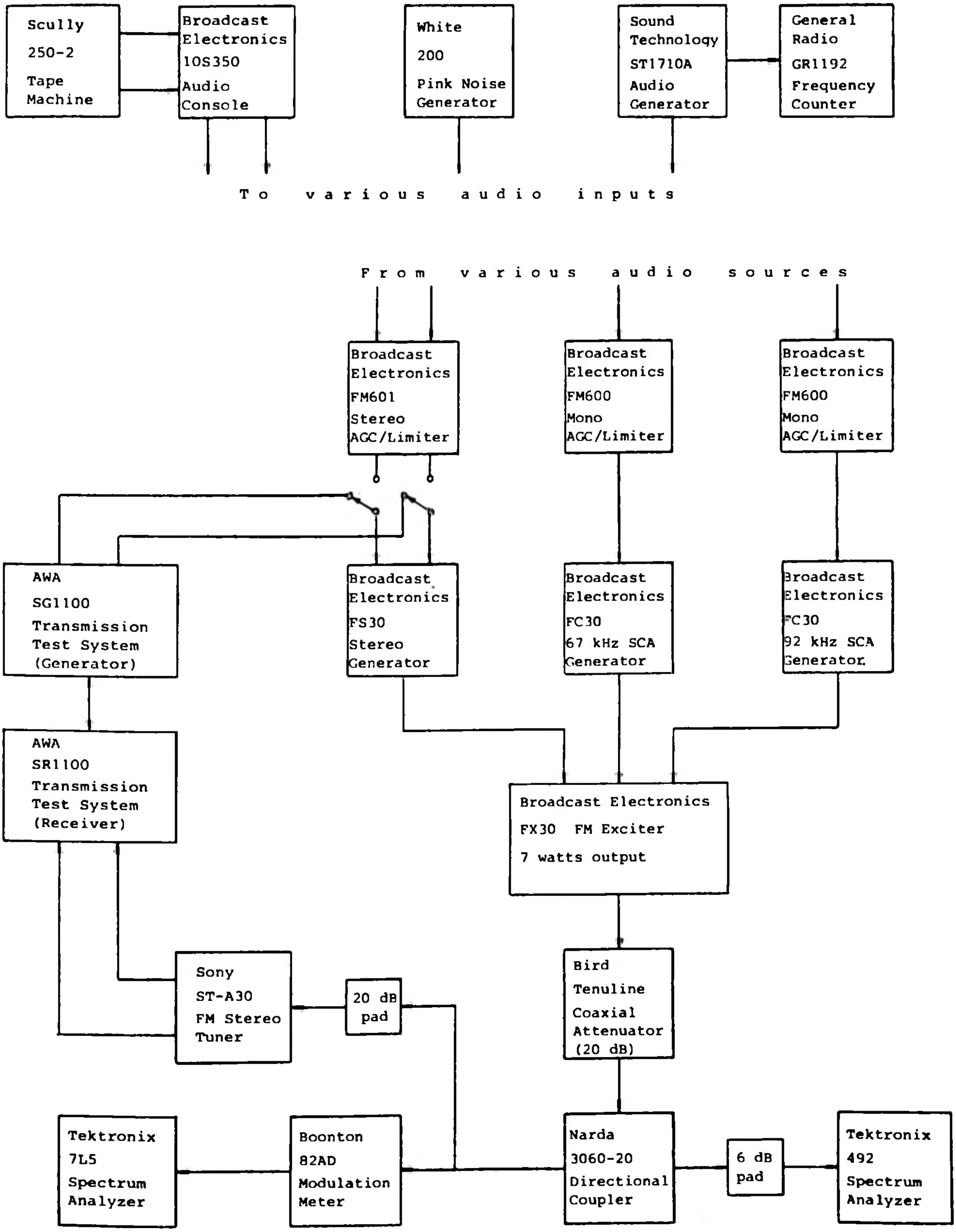


Figure 1

Block Diagram - Measurement Equipment Configuration

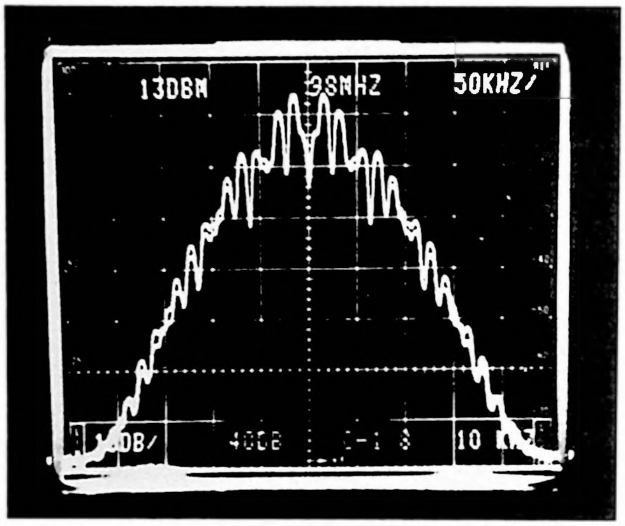


Figure 2a

Overlay of RF Spectra
View A and View B

View A: 91% stereo (15 kHz left ch.)
9% pilot
100%

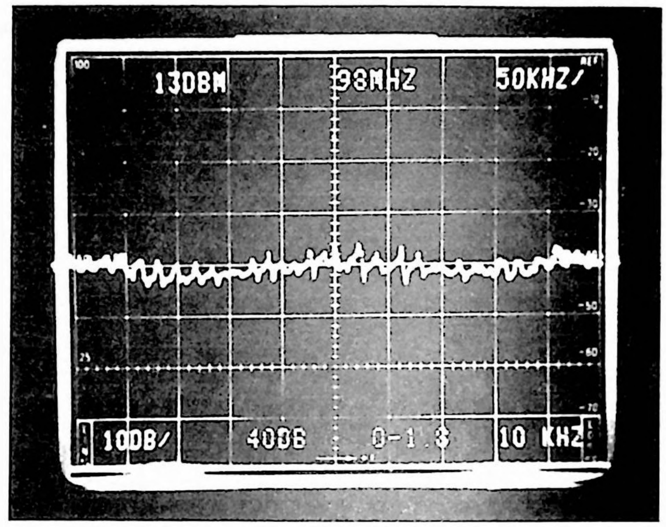


Figure 2b

Difference in RF Spectra
View B minus View A

View B: 82% stereo (15 kHz left ch.)
9% pilot
9% 67 kHz SCA
100%

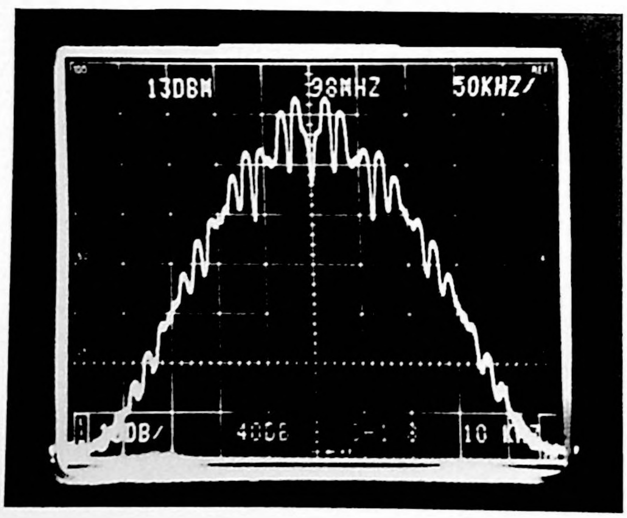


Figure 3a

Overlay of RF Spectra
View A and View B

View A: 91% stereo (15 kHz left ch.)
9% pilot
100%

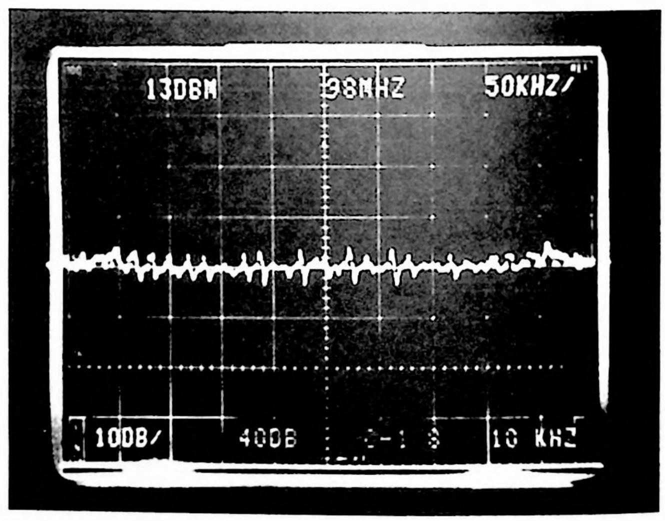


Figure 3b

Difference in RF Spectra
View B minus View A

View B: 87% stereo (15 kHz left ch.)
9% pilot
9% 67 kHz SCA
105%

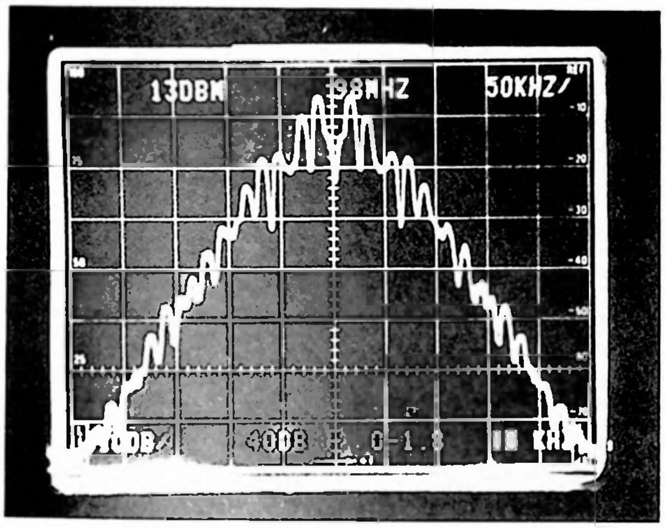


Figure 4a

Overlay of RF Spectra
View A and View B

View A: 82% stereo (15 kHz left ch.)
9% pilot
9% 92 kHz SCA
100%

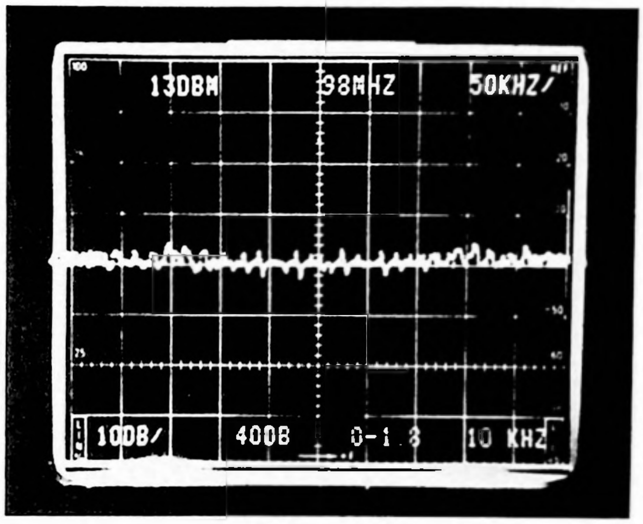


Figure 4b

Difference in RF Spectra
View B minus View A

View B: 87% stereo (15 kHz left ch.)
9% pilot
9% 92 kHz SCA
105%

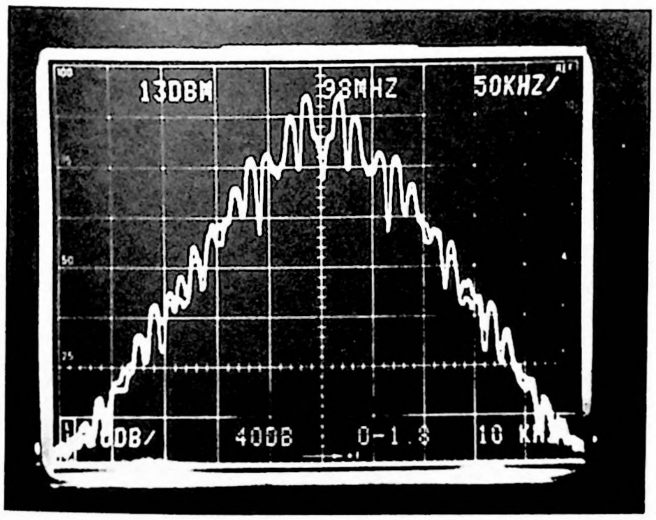


Figure 5a

Overlay of RF Spectra
View A and View B

View A: 82% stereo (15 kHz left ch.)
9% pilot
9% 92 kHz SCA
100%

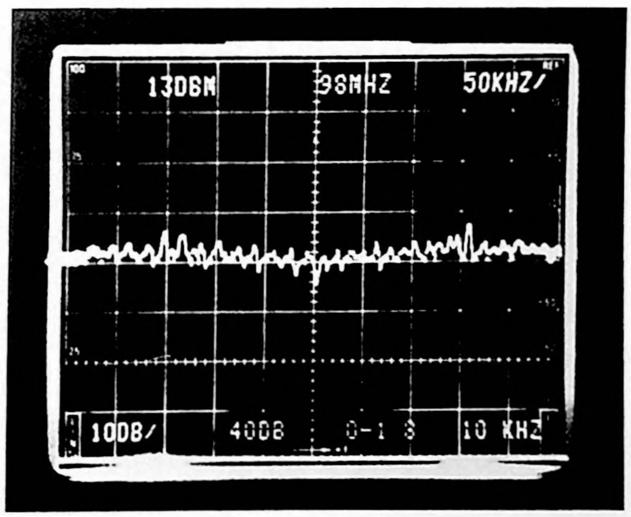


Figure 5b

Difference in RF Spectra
View B minus View A

View B: 82% stereo (15 kHz left ch.)
9% pilot
9% 67 kHz SCA
9% 92 kHz SCA
109%

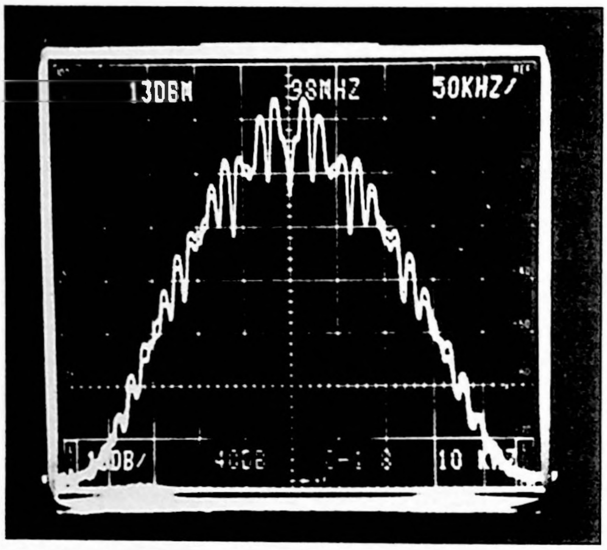


Figure 2a

Overlay of RF Spectra
View A and View B

View A: 91% stereo (15 kHz left ch.)
9% pilot
100%

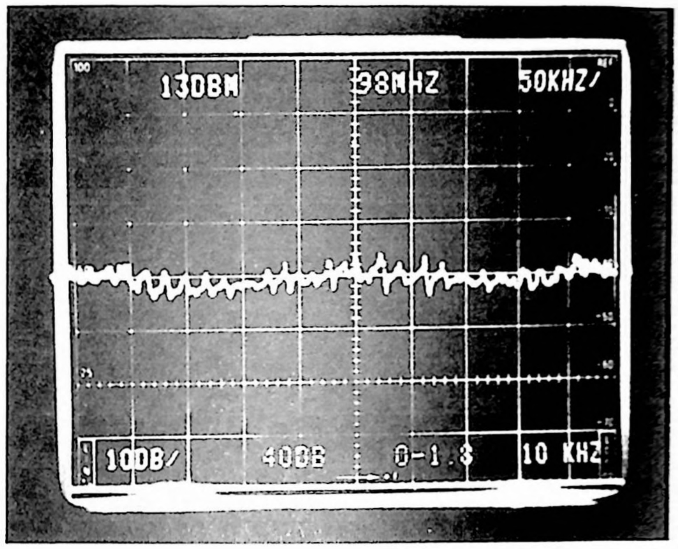


Figure 2b

Difference in RF Spectra
View B minus View A

View B: 82% stereo (15 kHz left ch.)
9% pilot
9% 67 kHz SCA
100%

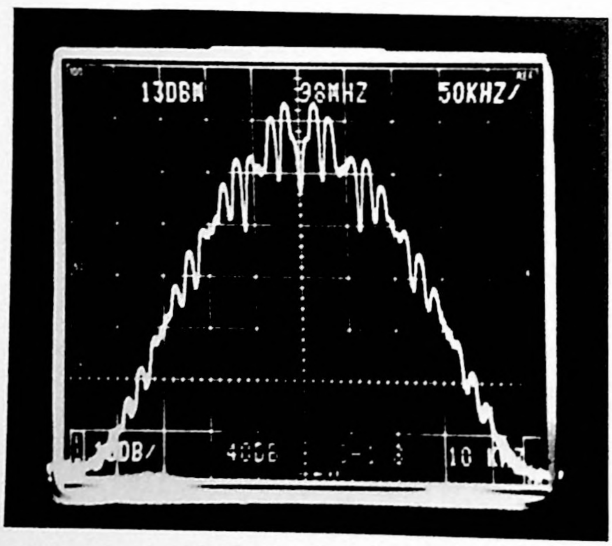


Figure 3a

Overlay of RF Spectra
View A and View B

View A: 91% stereo (15 kHz left ch.)
9% pilot
100%

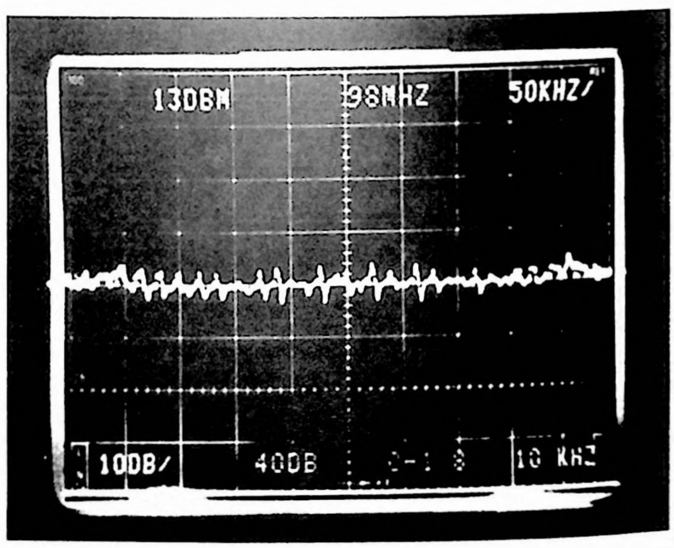


Figure 3b

Difference in RF Spectra
View B minus View A

View B: 87% stereo (15 kHz left ch.)
9% pilot
9% 67 kHz SCA
105%

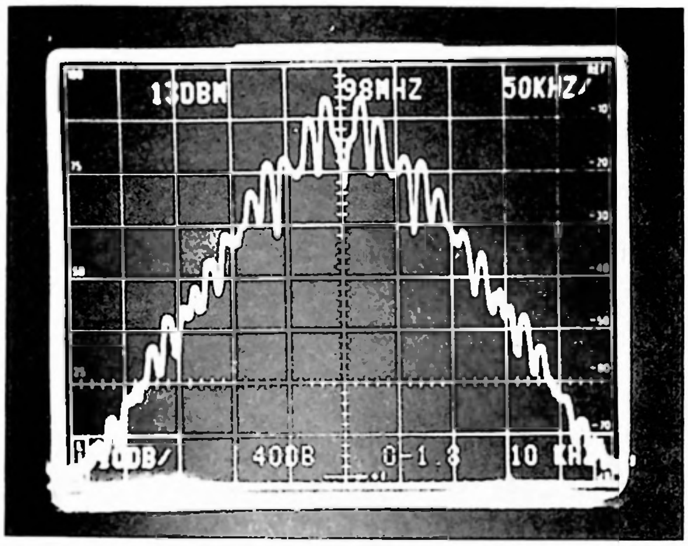


Figure 4a

Overlay of RF Spectra
View A and View B

View A: 82% stereo (15 kHz left ch.)
9% pilot
9% 92 kHz SCA
100%

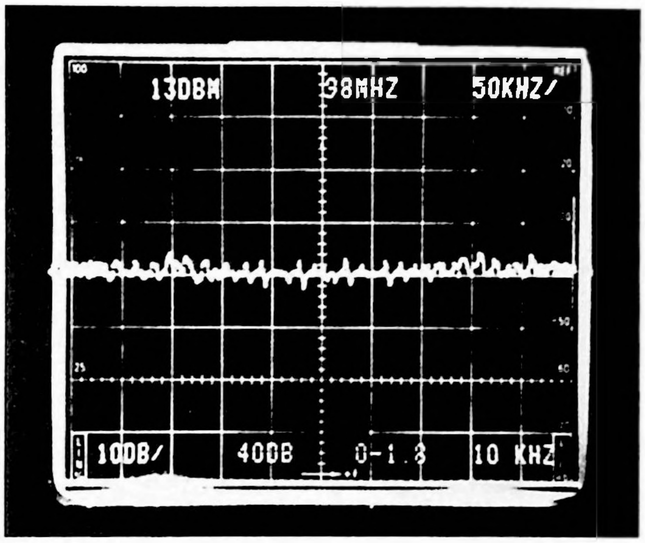


Figure 4b

Difference in RF Spectra
View B minus View A

View B: 87% stereo (15 kHz left ch.)
9% pilot
9% 92 kHz SCA
105%

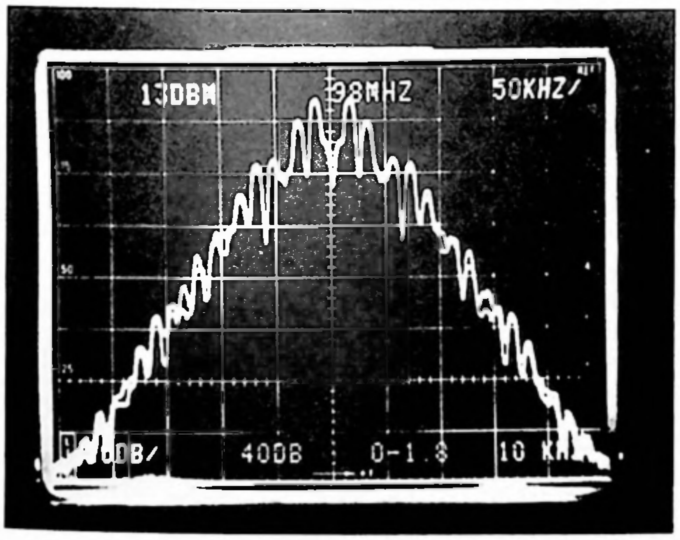


Figure 5a

Overlay of RF Spectra
View A and View B

View A: 82% stereo (15 kHz left ch.)
9% pilot
9% 92 kHz SCA
100%

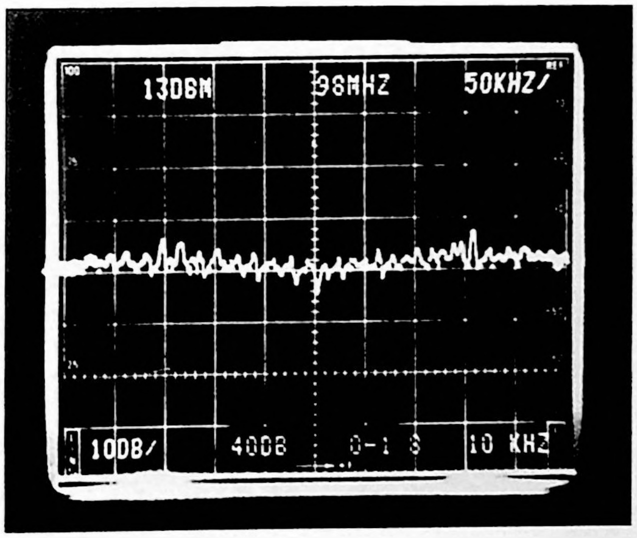


Figure 5b

Difference in RF Spectra
View B minus View A

View B: 82% stereo (15 kHz left ch.)
9% pilot
9% 67 kHz SCA
9% 92 kHz SCA
109%

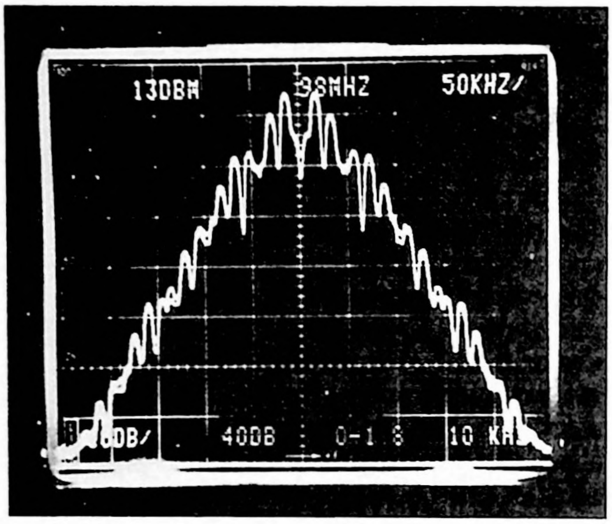


Figure 6a

Overlay of RF Spectra
View A and View B

View A: 87% stereo (15 kHz left ch.)
9% pilot
9% 92 kHz SCA
105%

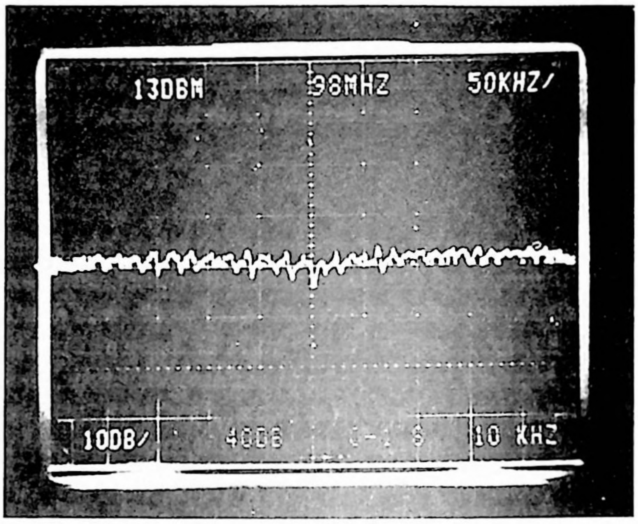


Figure 6b

Difference in RF Spectra
View B minus View A

View B: 87% stereo (15 kHz left ch.)
9% pilot
9% 67 kHz SCA
9% 92 kHz SCA
114%

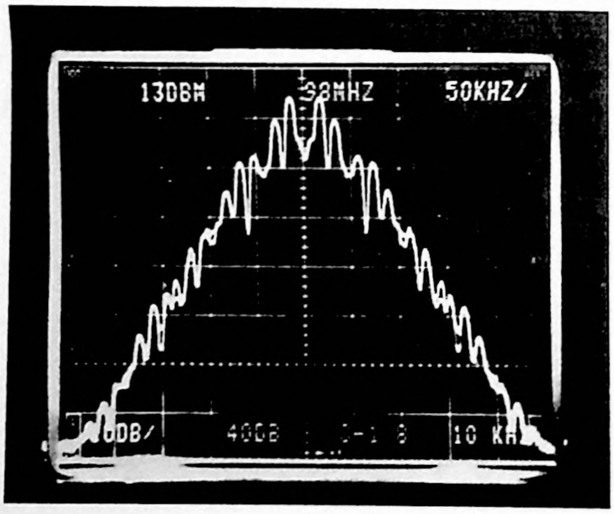


Figure 7a

Overlay of RF Spectra
View A and View B

View A: 82% stereo (15 kHz left ch.)
9% pilot
9% 92 kHz SCA
100%

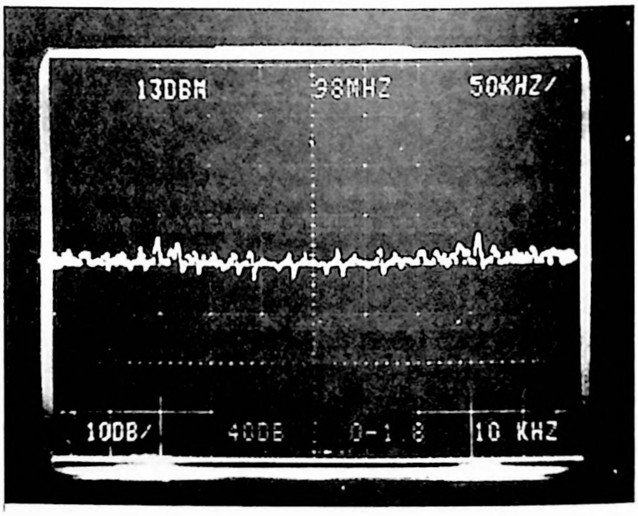


Figure 7b

Difference in RF Spectra
View B minus View A

View B: 87% stereo (15 kHz left ch.)
9% pilot
9% 67 kHz SCA
9% 92 kHz SCA
114%

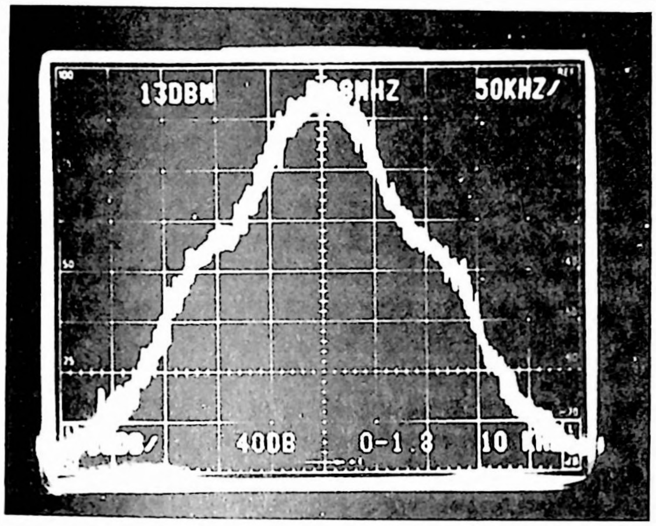


Figure 8a

Overlay of RF Spectra
View A and View B

View A: 82% stereo (program material)
9% pilot
9% 92 kHz SCA
100%

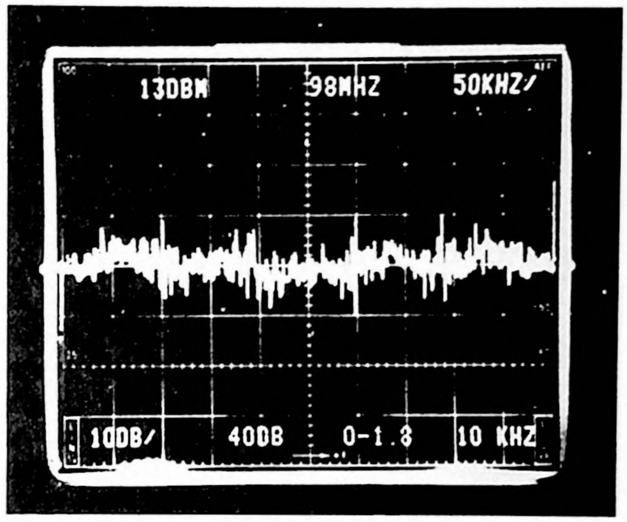


Figure 8b

Difference in RF Spectra
View B minus View A

View B: 87% stereo (program material)
9% pilot
9% 67 kHz SCA
9% 92 kHz SCA
114%

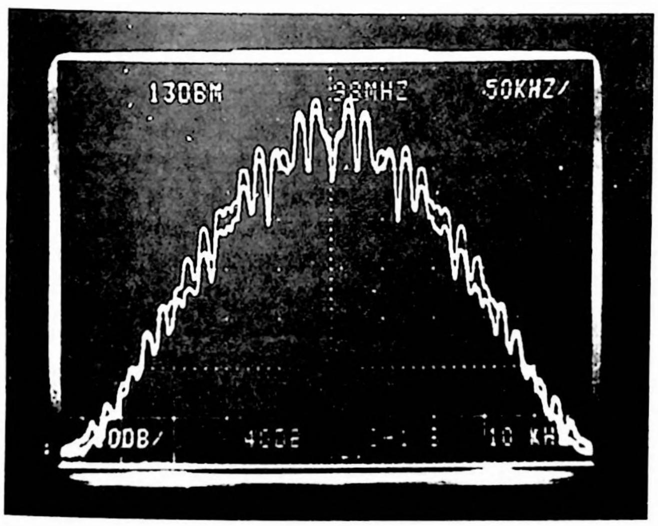


Figure 9a

Overlay of RF Spectra
View A and View B

View A: 87% stereo (15 kHz left ch.)
9% pilot
9% 67 kHz SCA
9% 92 kHz SCA
114%

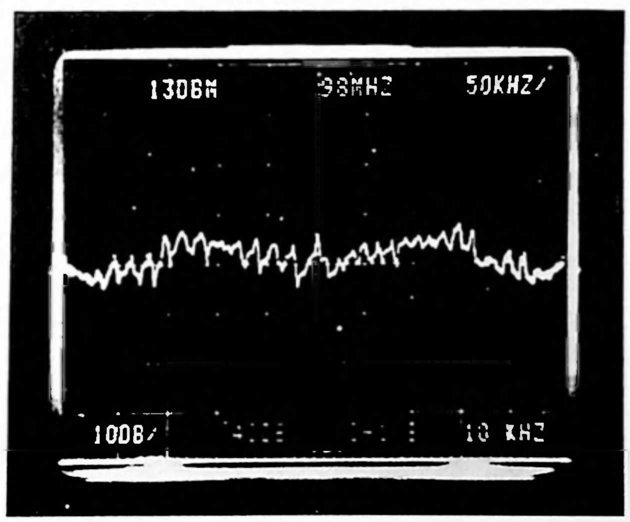


Figure 9b

Difference in RF Spectra
View B minus View A

View B: 105% stereo (15 kHz left ch.)
9% pilot
114%

TELEVISION MULTICHANNEL SOUND TESTS - CHICAGO 83

JAMES J. GIBSON
RCA LABORATORIES, PRINCETON, NJ

JAMES ZOOK
TEKTRONIX, INC., BEAVERTON, OREGON

Background and Objectives of the Tests

Under the sponsorship of the Consumer Electronics Group (CEG) of the Electronics Industry Association (EIA) and with the support of the National Association of Broadcasters (NAB) and the Joint Council of Inter-Society Coordination (JCIC), the Multichannel Sound (MCS) Subcommittee of the EIA Broadcast Television Systems Committee (BTS) was formed in 1979. The objective of the MCS subcommittee is to develop a technical record on which an industry recommendation for a single set of technical standards for the transmission of television multichannel sound can be based.

In August 1982 the MCS subcommittee presented to the BTS committee a 1000-page report in two volumes* containing the results of extensive tests and analyses on three proposed systems. The consensus of the BTS committee was that additional testing and analysis would be essential to evaluate certain critical technical questions relating to the proposed standards.

First, to optimize their systems, two of the proponents revised some system parameters subsequent to the testing program. Second, further tests were deemed necessary to evaluate potential difficulties imposed on cable television systems. Third, the original approach of selecting a companding system subsequent to the selection of a transmission system was deemed unrealistic and that testing of companding systems with realistic signal impairments would be necessary, primarily to evaluate coverage in fringe areas and cable television environments. Finally, a study of the test data indicated that the result of some of the tests intended to evaluate system performance appeared to have been obscured by the hardware design of the TV stereo receivers supplied by the proponents.

* "Multichannel Television Sound: The Basis for Selection of a Single Standard", by Electronic Industries Association's BTS Committee, published by the National Association of Broadcasters, Vol. I, July 16, 1982, Vol. II, August 6, 1982.

The objective of the continued test program is to provide the data needed to resolve these remaining technical uncertainties so that a single standard can be recommended with confidence by the BTS committee.

Tests on proposed transmission systems are currently conducted by one EIA-MCS task force (B/D) at the Matsushita Industrial Company in Chicago while tests on proposed companding systems will be initiated shortly by another task force (G) at the CBS Technology Center, Stamford, Conn. All the tests are scheduled to be completed during the summer of 1983.

This paper reports primarily on the transmission system tests in Chicago.

Review of Proposed Systems

There are three proponents of transmission systems:

- Electronic Industries Association of Japan ("EIAJ")
- Telesonics Systems, Inc. ("Telesonics")
- Zenith Radio Corporation ("Zenith")

There are also four proponents of companding systems:

- CBS, Inc. ("CX")
- dbx, Inc. ("dbx")
- Dolby Laboratories ("Dolby")
- Straight Wire Audio (SWA) under license from AEG-Telefunken ("SWA-HICOM")

Common aspects of the proposed transmission systems are:

- Supersonic subcarriers are added to the monophonic audio signal for FM multiplex transmission on the main sound carrier. One subcarrier carries a stereo difference signal, a second subcarrier carries a second audio program (SAP), and a third subcarrier is intended for Telemetry or non-public voice grade communication.

- No change in sound carrier power standards is proposed.

- For compatibility with existing receivers the monophonic standards are unchanged, including the 25 kHz peak deviation ("injection") by the monophonic signal.

- For stereo, the compatible monophonic channel carries a "Left plus Right" signal (L+R) while a subchannel carries the difference signal "Left minus Right" (L-R).

- A second audio program (SAP) is carried by an FM subcarrier above the stereo difference subcarrier. The SAP injection is 15 kHz and the deviation of the SAP subcarrier is 10 kHz.

- While no companding is proposed for the compatible monophonic (L+R) signal, the (L-R) signal and the second audio program signal (SAP) are compressed to improve the signal to background noise ratios and require corresponding expansion in the receiver.

• A non-public FM subchannel for Telemetry or voice quality audio is proposed with an injection of 3 kHz, a deviation of 3 kHz, and an audio bandwidth of 3.4 kHz.

The most important differences between the proposed systems are the modulation and the carrier frequency of the stereo difference signal:

• EIAJ carries the stereo difference signal (L-R) by frequency modulation of a subcarrier at $2f_H \cong 31.5$ kHz (the injection is 20 kHz and the deviation of the subcarrier is 10 kHz).

• Telesonics and Zenith carry the stereo difference signal (L-R) on an AM subcarrier (suppressed carrier). While the peak injections by the (L+R) audio signal are 25 kHz and 50 kHz by the (L-R) subchannel, the total injection by the combination of the monophonic channel, which carries (L+R) and the subchannel which carries 2(L-R), is limited to 50 kHz. As in FM-radio broadcasting the stereo difference signal is detected by synchronous detection using the phase of a pilot tone at half the subcarrier frequency as a reference.

• The subcarrier frequency of the Telesonics system is at $2.5f_H \cong 39.8$ kHz while in the Zenith system it is at $2f_H \cong 31.5$ kHz.

• The total peak deviation has been increased from the present 25 kHz to 65 kHz in the EIAJ-system, to 68.8 kHz in the Telesonics system, and to 75.5 kHz in the Zenith system.

Set-Up for Testing MCS Transmission Systems

The MCS subcommittee decided that all the remaining tests could best be made on a common, well-specified, closed circuit test bed, in which controlled transmission impairment could be injected.

Since the multichannel sound information in all the proposed systems is conveyed by multiplexed signals on the baseband of the aural carrier, it was judged that the interface between the test bed and the proponents' systems should be at the aural baseband. In this approach the proponents' MCS encoder feeds the aural baseband input of a common high quality television transmitter and the proponents' decoder is fed by the aural baseband output of a high quality television receiver.

Controlled transmission impairments are inserted in the r.f. path between the output of the transmitter and the input of the receiver. This approach assures that a comparative evaluation of the proposed systems will not be obscured by variations in receiver hardware design.

For the compatibility tests the r.f. signal is also distributed to the inputs of a number of monophonic receivers which are representative of receivers expected to be in the field a few years from now. A second visual transmitter is required to explore potential interference in the picture of the upper adjacent channel.

Figure 1 is a simplified block diagram of the test bed emphasizing the test philosophy of interfacing with the proponents' MCS systems at the aural baseband level.

The key components of the test bed are:

- A channel 3 visual modulator consisting of a Harris visual exciter (Model 994-7861) with adjustable incidental phase modulation and means for notch filter equalization.

- A channel 3 aural modulator consisting of a Boonton signal generator (Model 103D) feeding the output stage of a Harris aural exciter (Model 994-8013).

- A channel 3 Notch Diplexer (Microcommunications Model 51531-3) to combine the visual and aural carriers.

- An aural monitor (McMartin TBM-3500B) calibrated for 100% modulation at a 75 kHz deviation and capable of handling a 200 kHz baseband bandwidth.

- A television monitor (Tektronix 1450).

- A 75 ohm r.f. cable distribution network feeding 20 receivers.

- Means for inserting a translator or cable television processing equipment in the distribution system.

- Means for inserting controlled amounts of random noise, impulse noise, and multipath (or rather bi-path) in the distribution system.

- A channel 4 visual modulator for subjective evaluation of potential interference of channel 3 MCS-sound into channel 4 picture.

- Sixteen "compatibility" receivers for subjective evaluations of picture quality and monophonic audio quality.

- Three "compatibility" receivers with audio output terminals for quantitative measurements of monophonic quality.

- A high quality certified television receiver for reception of the aural baseband carrying the multichannel sound signals to be delivered to the proponent's decoder. This television receiver consists of a modified Tektronix 1450 monitor preceded by a low noise r.f. amplifier intended to operate at an input level of 10 dBmV (-39 dBm) with virtually noise-free aural baseband reception. The modifications of the Tektronix 1450 for standard (Nyquist) intercarrier reception, "quasi-parallel" intercarrier reception, and separate sound reception with sufficient bandwidth to accommodate all proposed MCS baseband signals are described in the next section.

Modifications and Performance of the Tektronix 1450-1 Demodulator

The Tektronix 1450-1 Demodulator was chosen for the system receiver. It is well suited for the task having carefully controlled frequency response and delay characteristics, wideband and narrowband surface acoustic wave (SAW) IF filters, and a quadrature output for use in incidental phase modulation (ICPM) display. In addition, it features a modular construction approach allowing easy access to each subsystem of the instrument. In addition, several unique features were needed:

1) An intercarrier sound down conversion method using symmetrical filtering about the visual carrier IF. This is referred to as "Quasi Parallel" intercarrier reception. This must be a separate filter from the usual Nyquist slope IF filter.

2) A split sound down conversion method must also be possible with a totally independent local oscillator (LO) replacing the LO generated from the visual IF.

3) A means of measuring ICPM without a receiver Nyquist slope filter causing AM to PM conversions.

4) A baseband aural output channel having a bandwidth of at least 100 kHz, flat within ±1/2 dB, and with minimal variations in delay.

The 1450-1 used as the test receiver was modified to provide the above four unique features. Figure 2 is a block diagram of the 1450-1 with the modified subsystems highlighted. To provide the first and third features the wideband Nyquist slope SAW filter of the IF section was replaced with an L-C bandpass filter. The new filter is symmetrical about the visual carrier IF with a 3 dB bandwidth of approximately 300 kHz. When this filter is user selected at the front panel, intercarrier sound down conversion occurs free of receiver-produced phase modulation components. Careful measurements of ICPM can be taken by using the quadrature output of the 1450-1.

The second feature listed above is provided by loop-through back panel BNCs carrying the split signal to the "Audio Source" switch. The split signal path can then be broken and a totally independent external signal used to convert the audio FM carrier to the standard 4.5 MHz. A Hewlett Packard HP8640 is used for the stable external source. All three methods of audio conversion can then be compared: intercarrier, split, and completely separate.

The fourth feature has been implemented by carefully readjusting the low pass filter in the audio section of an existing wideband audio option of the 1450-1, the option 2B.

With these changes implemented, the 1450-1 provides a quality demodulator with the flexibility needed from the systems test receiver.

Tests of Transmission Systems (1983)

A large number of the earlier tests are still applicable to currently proposed systems. For example, the good correlation obtained between measured and calculated signal-to-noise ratios indicates that no further random noise tests are required. The ratio of the peak video carrier power to the power of the aural carrier is 10 dB for broadcast tests and 15 dB for cable television tests. All tests will be done at a receiver input level of 10 dBmV (3.16 mV over 75 ohms). At this level front-end receiver noise will not mask other disturbances in any of the MCS subchannels (except possibly in the telemetry channel which is not tested for performance).

Some routine tests to certify the proper operation of the combination of the proponents' systems and the test bed are required. They include noise floor, stereo separation, and audio amplitude response vs. frequency.

An important test is the total distortion vs. a single tone sweep. For a left-only tone and with the SAP channel in operation, this is also a test of crosstalk for the case when the MCS baseband is fully activated. This test is consequently also a compatibility test. The single tone distortion tests will also be made for a typical case of multipath transmission (approximately a one-mile delayed path of 10% field strength and a worst case phase which is 90^o according to previous tests). In another compatibility test the crosstalk into a quiet main channel is determined when the SAP channel is active and the stereo subchannel is fully activated by a single tone sweep. This test is also made under multipath conditions.

In an intercarrier reception mode ("quasi-parallel" or "Nyquist") incidental phase modulation (ICPM) of the video carrier is transferred to the audio carrier in the intercarrier mixing process. This causes disturbances on the audio baseband which tend to look like subcarriers at multiples of the horizontal sync frequency f_H (15.75 kHz) modulated by multiples of the vertical sync frequency (60 Hz). Just like random noise in FM reception, the level of this disturbance tends to increase in proportion to the baseband frequency. The resulting background audio noise is referred to as buzz. This buzz tends to be concentrated at low frequencies in the Zenith system, at around 7.87 kHz in the Telesonics system, while in the EIAJ system and in the SAP channel it is spread over the entire audio band. In FM subchannels such as the SAP or the EIAJ stereo difference subchannels, a multiplicative disturbance referred to as buzz beat also results. The frequency of the buzz beat tends to be equal to and its amplitude proportional to the instantaneous frequency deviation of the FM subcarrier.

The buzz level is measured as a variation in the level of the noise floor with variations in the degree of ICPM, which is adjustable in the visual modulator. Buzz beat is detected in "total distortion" measurements and is subjectively evaluated in the companding tests. Buzz and buzz beat caused by the Nyquist slope in ordinary intercarrier reception is also measured for zero degree of transmitted ICPM.

Common mode phase modulation (CMPM), which can be generated by translators and cable television set-top converters, may also cause background noise particularly in a "separate sound" reception mode. Thus, while separate sound receivers are quite immune to ICPM, they are substantially more vulnerable to CMPM than intercarrier receivers. The effects of "typical" common mode phase modulation are determined by noise floor measurements.

The effects of impulse noise injected in the r.f. distribution system are recorded for subjective evaluation in the companding tests.

Subjective evaluations of sound quality and interference in the adjacent channel picture are made by several groups of observers on the 16 compatibility receivers. Tape recordings are also made of audio program material conveyed over the test bed with injection of various types of transmission impairments.

In addition to the planned test by the EIA-MCS subcommittee, the test bed will be used for the exploration of effects of cable television equipment. The cable television tests are discussed in a subsequent paper by Alex Best.

Interaction with Tests of Companding Systems

Tests on the companding systems are discussed in this paper only insofar as these tests interact with tests and analyses of the transmission systems.

The purpose of companding is to improve the subjective quality of audio signals degraded by transmission impairments. Comparative subjective tests of four companding systems are conducted over an audio signal test bed at the CBS Technology Center in Stamford, Connecticut. Figure 3 is a simplified block diagram of the test bed showing that only the (L-R) signal and the SAP signal are subject to companding. Figure 3 also shows that the transmission impairments are injected and specified in the audio frequency band. This is in contrast to the injection and specifications of impairments at r.f. frequencies in the transmission test bed as illustrated in Figure 1. Thus a conversion of the effects of the r.f. impairments to audio impairments is required.

The companding systems will be tested for four types of impairment: random noise, impulse noise, buzz, and buzz beat. The magnitude and spectral distribution of random noise can be determined analytically with great accuracy. Based on this analysis random noise is generated locally by noise generators. Buzz beat, which is a multiplicative noise, must also be generated locally by a "buzz beat" generator. The present plan is to record buzz and impulse noise in the output of the decoders in the transmission system test bed for subsequent injection in the companding system test bed.

Peak flashers are used to monitor the peak signal levels which are constrained by the transmission systems. The monophonic signal is constrained by a peak deviation of the (L+R) signal to 25 kHz and the SAP signal to a peak deviation of the SAP subcarrier of 10 kHz. In the EIAJ system the (L-R) subcarrier is also constrained to a peak deviation to 10 kHz, while in the Zenith and Telesonics systems the signals (L+R) ± 2(L-R) are constrained to peak deviations of 50 kHz.

The levels of the injected impairments are adjustable and measurable with various instrumentations.

Conclusions

A test bed for testing television multichannel transmission sound systems has been implemented by the EIA-TVMCS subcommittee at the Matsushita Industrial Company in Chicago and is currently used to complete tests on three proposed MCS systems. The basic approach is to interface with the proponents' systems at the baseband level by using a common transmitter, a common receiver, and inject controlled transmission impairments in the r.f. path.

Acknowledgments

The continued MCS tests in Chicago have received broad support from the industry. The contributions to the test plan by members of the EIA-TVMCS subcommittee and by the system proponents are greatly appreciated. A special thanks goes to our host, the Matsushita Industrial Company, and to EIA consultants Carl Olson and John Landeck who have implemented the transmission test bed and currently carry out the tests.

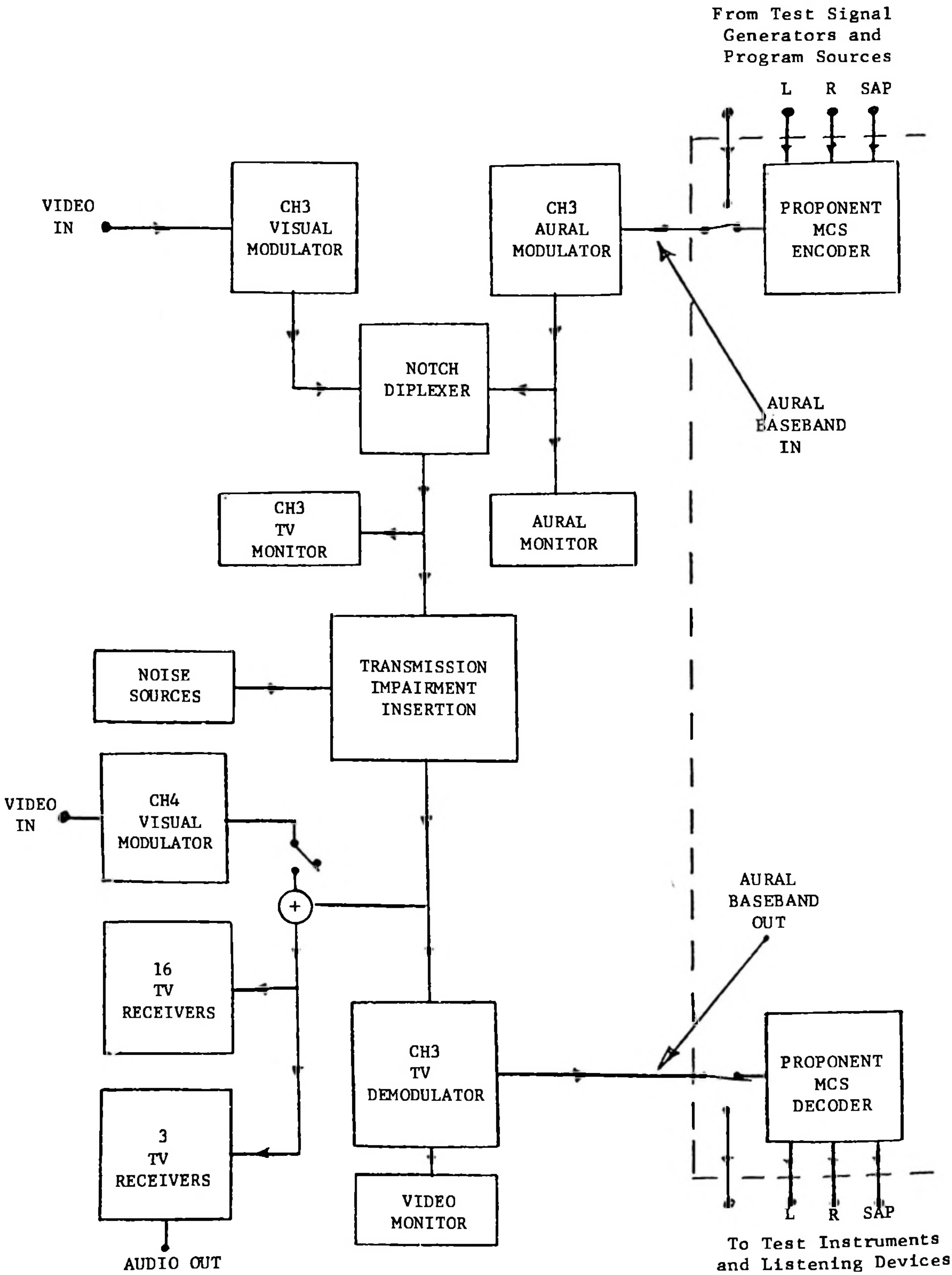


Figure 1: Block Diagram of MCS Test Bed

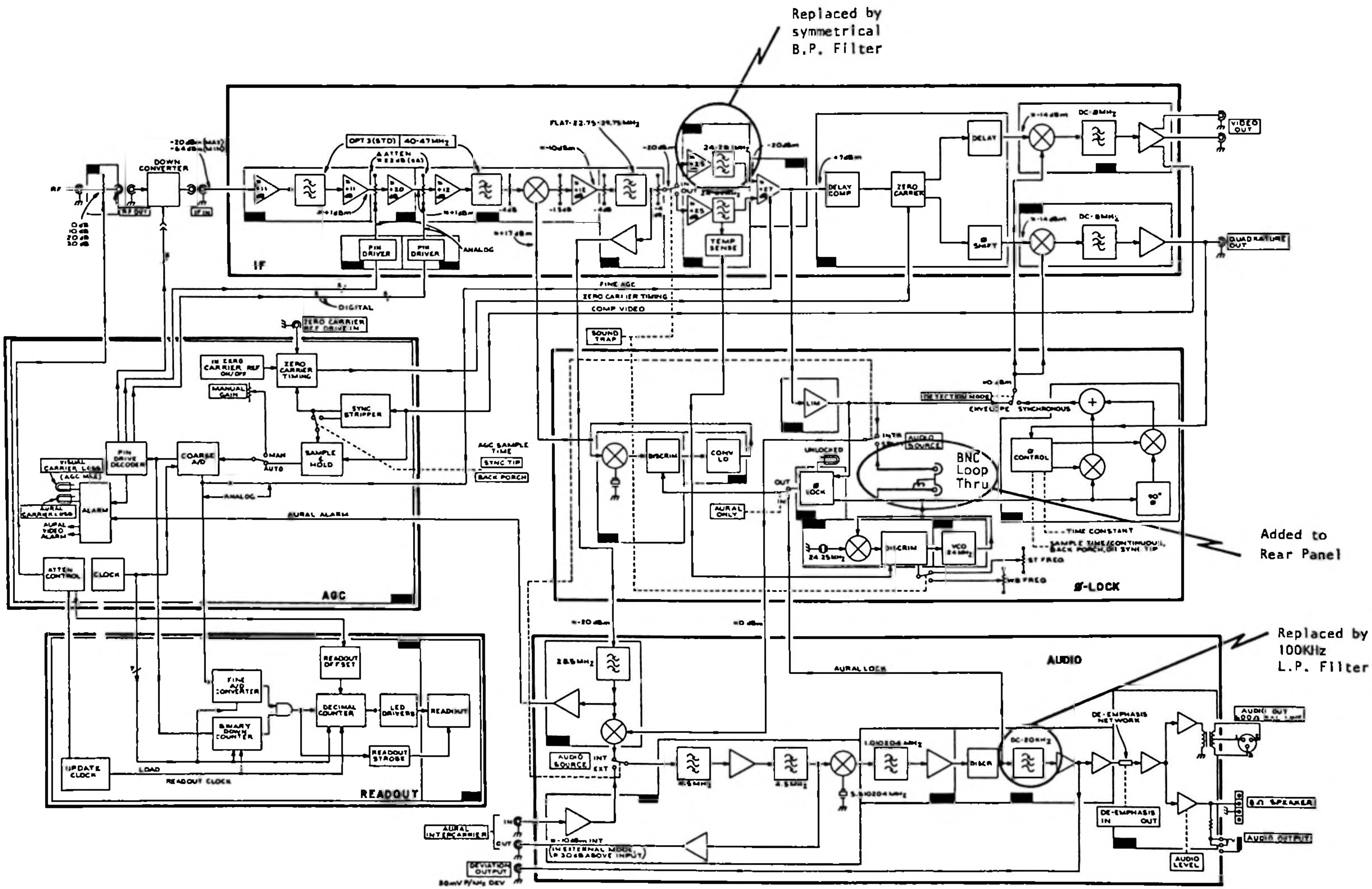


Figure 2: 1450-1 Block Diagram

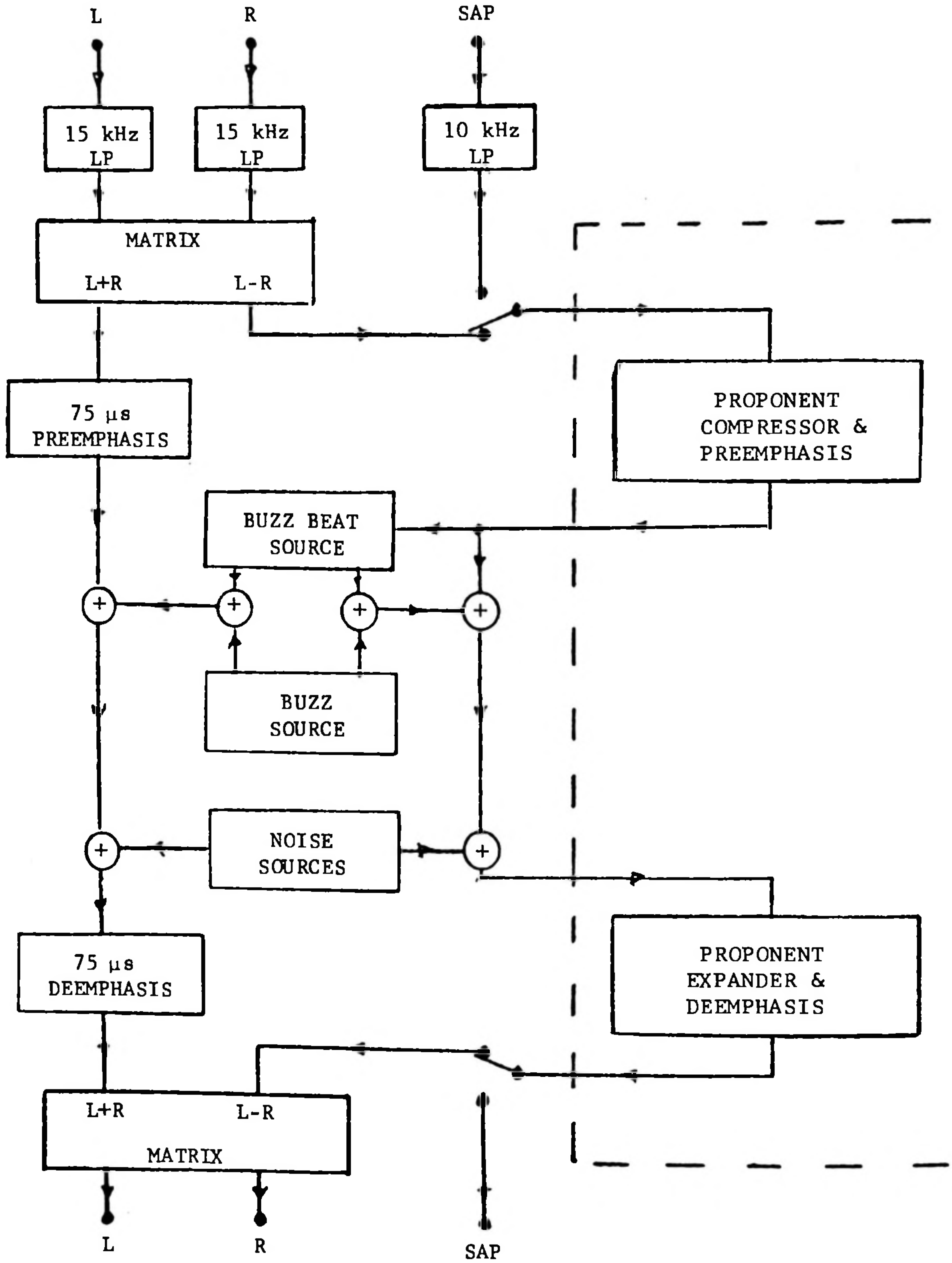


Figure 3: Test Bed For Companding Systems

INTERCARRIER BUZZ IN TELEVISION RECEIVERS

CARL G. EILERS
ZENITH RADIO CORPORATION
GLENVIEW, IL 60025

I. Introduction

Intercarrier buzz is a phenomenon that receiver designers have contended with for years. Recent interest in improved (monophonic) television sound, in stereo and multichannel sound and in other subcarrier uses has stimulated work in understanding the sources of buzz. This paper points to cures together with the compromises and tradeoffs involved.

- There are many sources of buzz in the television system:

1) ICPM of the visual transmitter.
2) Inadequate (or no) filtering of the transmitted visual sidebands in the vicinity of the aural channel.
3) Video coupling to the audio through common power supplies used in some television transmitters.
4) Inadequate filtering of video signals in microwave equipment (used for networking, or STL applications) for the protection of the audio subcarriers. Or, additionally, suboptimum performance of this same equipment in the microwave FM process, i.e., differential phase/gain, giving rise to buzz.
5) Common visual and aural processing in translator broadcast stations, giving rise to buzz (especially bad for split sound reception). Extra phase noise (and buzz) generated in the heterodyne oscillators used in such equipment is also especially bad for split sound receivers.
6) Similar common visual and aural processing used in cable headend and subscriber converter equipment. Comments in 5) above apply also to this case.
7) Lowered aural carrier level in relation to the visual carrier level for signals delivered by CATV systems to television receivers having some nonlinearities in tuner and/or i.f. amplifiers will generally give rise to additional buzz compared to broadcast reception by the same receiver. Nonlinearities in the tuner will affect split sound reception as well as intercarrier reception.
8) Split sound as well as intercarrier television receivers will buzz on reception of monophonic signals delivered by video cassette players, video disc players, TV games, and similar sources. Intercarrier receivers, however, are more forgiving in this case.

Buzz results when video related, phase modulated components of the visual carrier are transferred to the carrier that transmits the sound. The vertical blank and sync components provide the characteristic buzz sound. In monophonic sound, especially when reproduced with limited low and high frequency response and moderate listening level, the buzz is, most of the time, barely noticeable and certainly acceptable. But when multichannel sound is attempted with subcarriers it is found that these elevated baseband frequencies are much more vulnerable to ICPM. In fact the property of FM by which thermal noise increases 6 dB for an octave frequency increase is also applicable to buzz interference. It thus becomes important to have quantitative data on the various causes of ICPM. There is a rather extensive literature available on ICPM as relating to chroma [1], [2] , [3], [4] , [5], but less is available concerning sound reproduction [6], [7], [12].

It is the intention of this paper to set the causes of ICPM in perspective, give quantitative results, to analyze current transmitter and receiver design practices and to indicate ways to improvement in the TV receiver.

II. Analysis of Intercarrier Buzz

A. Sources of Buzz

The generation of buzz is principally a 2-stage process, first of ICPM of the visual carrier and secondly - by a distinct process - transfer of this ICPM to the intercarrier.

Here is a list of ICPM sources:

1. Level-dependent phase shift
   a. Parametric effects
   b. Transmit time effects
2. AM-PM Conversion
   a. Unequal sideband amplitude ("Nyquist ICPM")
   b. Unsymmetric sideband phase
   c. Straight FM (contamination of local oscillator)
3. Miscellaneous
   a. Overmodulation on white
   b. Reverse mixer feedthrough
   c. 3-Tone intermodulation (?)
   d. CATV equipment

B. Transfer of ICPM to the Intercarrier; Detection

The simplest type of second detector which is still extensively used is the envelope detector. The visual carrier acts as local oscillator for the suitably attenuated aural carrier and this produces the Intercarrier as "beat" product. If the visual carrier is amplitude modulated only and if this amplitude modulation is transferred to the intercarrier it will be eliminated by the limiter. Any angle modulation which is only on the visual carrier, however, is fully transferred to the intercarrier. The FM sound detector detects these video signal-related angle modulations. In case they are in the audio frequency range or in the composite baseband range of frequencies for multichannel sound they become audible. The vertical sync and blank components cause the characteristic buzz sound.

When a synchronous second detector is used ICPM transfers by a somewhat different process, although the result may be the same. Assume that the Reference Channel (into which the IF signal is branched in order to derive a local carrier) has a bandwidth which is wider than twice the composite baseband width of the multichannel audio. As a result the locally generated carrier still contains the ICPM. Thus ICPM is transferred to the intercarrier.

Next assume a synchronous detector with a reference channel only a few kHz wide (such as in a TV broadcast station demodulator). All ICPM with a rate faster than a few kHz is now eliminated from the local carrier and thus will not transfer to the intercarrier.

There is yet another process of interest in the second detector. Assume that both visual and aural carriers have acquired equal amounts of ICPM. This could, for example, happen in a tuner that has an insufficiently bypassed AFC voltage. Consequently video products are applied to the frequency control in the tuner local oscillator. This is a straight FM modulation process and it acts on both visual and aural carriers. The interesting result in an envelope type second detector is that the ICPM is eliminated from the intercarrier since it is the result of differencing of visual and aural carriers. A synchronous detector with wide band reference channel acts similarly on the intercarrier. When the reference channel is narrow, however, the local oscillator has no significant ICPM while the aural carrier does have. The result is an intercarrier <u>with</u> ICPM. Note that the envelope detector and the synchronous detector with sufficiently wide reference channel process ICPM in a like manner in contrast to the way in which a synchronous detector with narrow reference channel processes ICPM.

In the early days of television, receivers had separate processing of visual and aural carriers. This required extra vacuum tubes and it made tuning much more critical,especially on UHF, so intercarrier detection was an improvement. But the split processing typically had no buzz!

They had no buzz because the aural carrier was received without ICPM, the tuner did not introduce any (no AFC in those days) and after the tuner both carriers were separated. The aural carrier had its own bandpass filter and no Nyquist slope introduced ICPM (see section 2. below). With present-day sophisticated tuning systems and with integrated circuits, split-sound processing is becoming attractive again provided that no tuner ICPM is introduced. This is taken up in section III.

1. Level Dependent Phase Shift; Transmitters

When in some stage or device the phase shift of a small signal varies with bias level (= luminance level) the small signal experiences differential phase distortion while the luminance modulated carrier experiences ICPM. This is a nonlinear effect. Such a nonlinearity causes harmonic distortion, intermodulation and differential gain distortion but not, per se, differential phase distortion. Differential phase distortion can best be explained by parametric effects and/or by transit time effects. The former include dependence of junction capacities, dynamic resistances, etc. on current and/or voltage. Reactive properties of such a circuit become dependent on bias level. Transit time variations, as take place, for example, in Klystron amplifiers cause similar phase effects. The consequences for the chroma signal are described by Behrend [1], Blair [2], and Kuroki [4], The transmitter is the source of the ICPM in these references. In U.S. Television broadcasting the aural carrier is processed separate from the visual carrier up to passive diplexer (except for internally diplexed transmitters for lowpower television or translator service).

Thus no direct transfer of ICPM to the aural carrier takes place in the transmitter, although similar parametric effects may take place in receiver tuner and IF. Mostly, though, the ICPM reaches the intercarrier via the second detector as described above.

Quantitative Analysis

On the basis of Behrend's measurements [1] the ICPM is assumed to vary in direct proportion to the carrier level with θ_o degrees per IRE unit. Figure 1 shows this graphically. Current transmitters appear to be able to hold the ICPM within 1^o over the 120 IRE units not including sync.

Figure 2 shows how the signal-to-buzz ratio due to this effect varies with modulating frequency. The curve marked "Parametric/Transit Time ICPM" is for a sine wave of 100 IRE units peak-peak and .01 degrees/IRE unit ICPM. Curves for other levels are parallel and can be found in the way indicated in Figure 2. Note that the SBR decreases 6 dB per octave. This is the result of the frequency independence of θ_o (which is justified for the range of a few dozen kHz around 45 MHz) combined with the triangular interference sensitivity of FM. When in a later section the Nyquist slope-caused ICPM is analyzed it is found that the ICPM angle itself increases 6 dB per octave increase in modulating frequency, resulting in 12 dB per octave buzz increase. (This is represented by the other curve in Figure 2.)

Sine wave modulation is easily treated but composite blank and sync combined with a constant grey level (including black and white) is more realistic. Signal-to-Buzz Ratio numbers for these signals are collected in Table I. The third entry is also found at 31.5 kHz in Figure 2.

The other entries are mainly for blank and composite blank and all are for constant white level.

The most interesting numbers are for composite blank. In the one case the vertical sidebands around the 2nd harmonic of horizontal are AM demodulated by $2f_h$ = 31.468 kHz to the audio band (59.1 dB) and the other case is for demodulation by $2.5f_h$ = 39.335 kHz (58.5 dB) of the upper vertical sidebands of the 2nd harmonic of horizontals and the lower vertical sidebands of the 3rd harmonic of horizontals. Note that both numbers include de-emphasis.

2. AM-PM Conversion; Nyquist Slope

   a. Under consideration here are strictly linear effects. When a double sideband AM modulated signal is passed through a passive network that results in unequal sideband amplitude or unsymmetrical sideband phase shift or both, the result is phase modulation. This is demonstrated by the well-known vector diagram of Figure 3 where the resultant vector "wobbles" around the unmodulated position at the modulation rate.

      Deutsch [8] has given expressions for $\theta = \theta(t)$ as well as for its time derivative, to which the FM detector output is proportional.

      Quantitative Analysis:

      Assuming a linearly decreasing Nyquist slope from 750 kHz below the carrier to 750 kHz above the carrier the ICPM is completely determined for a given modulating frequency and depth of modulation. For 100 IRE units peak-to-peak sine wave the resulting signal-to-buzz ratio is graphed in Figure 2. For less than 100 IRE units modulation the graphs are parallel to the given one and shifted upward by an amount of D dB which is found in Figure 4.

   b. The case of unsymmetrical sideband phase shift and sine wave modulation of 100 IRE units peak-to-peak is plotted in Figure 5. The SBR is plotted as a function of the angle of unsymmetry for a frequency of 31.5 kHz (actually the result of this ICPM is not buzz but a modulated tone of 31.5 kHz). Note that a phase unsymmetry of 0.2 degrees causes 45.3 dB signal-to-interference ratio. The sideband amplitudes are assumed equal.

   c. The results of some calculations for composite blank are collected in Table II. Unlike for the parametric ICPM case with linear phase slope the SBR is here not proportional to the peak-to-peak modulation. Figure 4 only holds for sine wave modulation.

      An important comparison for the viability of multichannel sound is between the composite blank data of Table I and Table II both demodulated to the audio band. Table I: 59.1 dB (1 degree total ICPM) and Table II: 44.2 dB (100 IRE units peak-to-peak modulation.

      The conclusion to be drawn from these numbers is that for a successful multichannel sound system it is important to eliminate the Nyquist ICPM. Once this is accomplished the black-to-white transmitter ICPM should be limited to 1 degree or preferably less.

d. Another source of ICPM is the tuner AFC to which reference was made in section II. B above. One of the objectives of TV receiver design is to have a fast acting AFC. This is desirable to allow a quick scan of the TV channels without having to noticeably wait for the channels to lock in. Faster action requires more bandwidth and this allows more video components to reach the tuner VCO. This causes direct FM modulation of the l.o. with subsequent transfer to visual and aural carrier alike. A steeper rolloff of the AFC filter is constrained by the stability requirement for the AFC feedback loop. Fortunately the described ICPM need not cause buzz if intercarrier detection is used with an envelope detector or a synchronous detector with a wide enough reference channel bandwidth. The split sound or parallel sound receiver type will have to contend with this ICPM as will the receiver with a synchronous detector with narrow reference channel. Microphonics and l.o. phase noise are also problems.

3. Miscellaneous ICPM Sources

a. Overmodulation

Video overmodulation generally causes severe buzz. When the visual carrier is modulated close to zero the 4.5 MHz "beat" from an envelope detector disappears. The FM detector loses capture and noise comes up. The carrier always returns at blanks which gives the noise the buzz character.

The synchronous detector with wideband reference channel, on visual overmodulation, loses reference and thus could cause a similar effect in the sound as the envelope detector, depending on the fly-wheel action of the 45.75 MHz oscillator.

In the case of separate processing of visual and aural carriers, overmodulation has no effect on the sound.

b. Reverse Mixer Feedthrough

With insufficient balancing and buffering the RF signal may reach the tuner local oscillator through the mixer. The AM may cause the instantaneous phase of the l.o. to vary. This constitutes ICPM and is, really, of the parametric kind. The oscillator Q is, generally, not high enough to eliminate any of this ICPM, just as in the case of AFC - introduced ICPM.

c. CATV Processing

Set-top heterodyne converters are used in most CATV installations. Many of the above described buzz phenomena are present in these devices when feeding split sound receivers.

C. Summary

Buzz is caused by video related incidental phase modulation of the aural carrier or intercarrier. This can be caused by transfer of ICPM from the visual carrier in a detector sensitive to this transfer. ICPM of the visual carrier can be caused by parametric or transit time effects in the transmitter or by unsymmetric sideband treatment anywhere but especially on the Nyquist slope. When aural and visual carriers are co-processed parametric effects can cause direct ICPM of the aural carrier. If identical ICPM is present on aural and visual carrier the conventional types of 2nd detector will eliminate this ICPM from the intercarrier. When a synchronous second detector has a narrow band reference channel (few kHz) the significant ICPM is removed from the regenerated carrier and this will produce an unadulterated intercarrier if the tuner does not contaminate the aural carrier.

III. Cures for Intercarrier Buzz

A. General Considerations

Until recently buzz has not been considered a significant problem. Add to this the complexity of the problem and it will not be a surprise that there has not been much done to improve the situation. Recently, however, the pressure for such solutions has been rising. There is increased interest in good quality monophonic sound; multichannel sound with its increased baseband width requires solutions; some Pay TV systems use subcarriers which should not result in degraded sound quality.

After a few short paragraphs on transmitter cures for buzz, there will follow descriptions of a number of alternate TV receiver circuit arrangements from a buzz point of view. The purpose is to provide insight into the interactions and tradeoffs involved. It could serve as the basis for further work.

B. Transmitter Cures for Buzz

Aural transmitters in the U.S. are separate from visual transmitters and thus are free of ICPM except when internally diplexed.

Visual transmitters require, in general, different correction methods for the high efficiency, high level circuits than for the low level circuits. For example, in a low level modulator the ICPM can often be cured by optimum adjustment of carrier injection levels and/or by feeding small amounts of quadrature carrier around the modulator.

High level circuits require pre-correction. Transmitter exciters are now becoming available with several levels of correction relative to visual modulation, including sync.

C. Receiver Cures for Buzz

1. The Split Sound Receiver

Also named: "Parallel Sound Receiver" or "Split-Carrier Sound Receiver"

This method (Figure 6) of sound processing was used in the early days of television before intercarrier detection was introduced. Only tuner ICPM can cause buzz because only in the tuner are visual and aural carriers co-processed. Video derived AFC can also cause ICPM. Signals delivered by internally diplexed transmitters or by set-top CATV converters will generally cause problems.

2. The Quasi-Split Sound Receiver

Also named: "Quasi Parallel Sound Receiver"

A more descriptive name would be "Split-Intercarrier Sound Receiver."

This system (Figure 7) also uses a separate sound channel but uses intercarrier sound detection and is thus insensitive to ICPM caused in the tuner. The Nyquist ICPM is eliminated by a specially designed IF filter that has symmetrical response around the visual carrier. The aural 2nd detector will not eliminate transmitter ICPM if it is of the envelope type or of the synchronous type with wide reference channel.

The latter type has a thermal noise advantage over the former. The visual carrier is contaminated by, among other, quadrature noise. The smaller the carrier (white signal, overmodulation) the larger the resulting phase angle. Thus in envelope detection the largest noise angle can be transferred to the intercarrier whereas in synchronous detection with constant local oscillator amplitude the smallest one is transferred.

If a narrowband reference channel were used for the synchronous detector local oscillator then transmitter ICPM problems would also be eliminated, to the degree that it falls outside of the reference channel. An additional advantage would be reduced phase noise transfer to the intercarrier especially at the higher audio baseband frequencies where it counts most. Unfortunately the narrowband requires extra tuning accuracy and such receivers will not work on most VCRs and TV games.

There have been reports of dual-output-SAW devices that can provide the conventional Nyquist-sloped output as well as the one indicated in Figure 7.

The circuits needed are not simpler than for the Split Sound Receiver but several advantages are gained. Insensitivity to tuner ICPM; the advantage of normal tuner tolerance as compared to the much tighter tolerance needed for the split receiver tuning. If transmitters that convert to stereo sound minimize their ICPM at the same time as new receivers become available, the sensitivity of quasi-split sound processing to transmitter ICPM no longer remains a problem.

3. Additional Improvement

Appropriately designed companding systems for noise reduction will also be effective on buzz if not excessive.

Conclusion

This paper has attempted to identify the sources of buzz and suggest cures, particularly in the TV receiver. For improved system performance: first the receiver Nyquist slope effect has to be eliminated, next the transmitter ICPM must be minimized and, finally, any combined visual/aural processing must be done in circuits with sufficiently low parametric nonlinearity.

One Final Conclusion

Buzz is a systems problem. It will take cooperation between transmitter manufacturers and operators, cable equipment manufacturers and cable system operations and receiver manufacturers to bring it down to acceptable levels - if not eliminate it.

Table I

TRANSMITTER ICPM (PARAMETRIC/TRANSIT TIME EFFECTS) TOTAL ICPM ANGLE = 1 DEGREE*				
MODULATING SIGNAL		RECEIVED SIGNAL	DE-EMPHASIS (?)	SIGNAL/BUZZ RATIO (Calculated, Referenced to 25 kHz Deviation) dB*
VERTICAL BLANK		MONO	NO	72.6
COMPOSITE BLANK		MONO	Y	76.8
31.5 kHz SINE WAVE		DIRECT	NO	39.2
HORIZONTAL BLANK				
	FUNDAMENTAL	DIRECT	NO	35.7
	2nd HARMONIC	DIRECT	NO	44.9
	3rd HARMONIC	DIRECT	NO	43.8
COMPOSITE BLANK				
	EIA-J	STEREO	Y	74.2
	TELESONICS - 2X	STEREO	Y	58.5
	ZENITH - 2X	STEREO	Y	59.1
COMPOSITE SYNC				
	ZENITH - 2X	STEREO	NO	71.9

*FOR 0.5 DEGREE ADD 6 dB, FOR 2 DEGREES DEDUCT 6 dB, ETC.

Table II

NYQUIST ICPM (AM/PM CONVERSION) MODULATING SIGNAL LEVEL = 100 IRE UNITS			
MODULATING SIDNAL	RECEIVED SIGNAL	DE-EMPHASIS	SIGNAL/BUZZ RATIO (Calculated, Referenced to 25 kHz Deviation) dB
VERTICAL BLANK	MONO	Y	76.1
COMPOSITE BLANK	MONO	Y	75.1
31.5 kHz SINE WAVE	DIRECT	N	23.9
HORIZONTAL BLANK			
FUNDAMENTAL	DIRECT	N	36.7
2nd HARMONIC	DIRECT	N	26.4
3rd HARMONIC	DIRECT	N	22.5
COMPOSITE BLANK			
EIA-J	STEREO	Y	57.6
TELESONICS - 2X	STEREO	Y	38.5
ZENITH - 2X	STEREO	Y	44.2

REFERENCES

[1] W. L. Behrend, "Effects of Incidental Phase Modulation of TV Transmitters, or Other Circuits, on TV Signals", IEEE Trans. on Broadcasting, VOL. BC-19, #3, Sept. 1973, pp 53-63.

[2] R. H. Blair, "Comments on [1]", IEEE Trans. on Broadcasting, VOL. BC-21, #2, June 1975, pp 38-39.

[3] S. Kuroki, "Comments on [2]", IEEE Trans. on Broadcasting, VOL. BC-22, #1, March 1976, pp 18-19.

[4] S, Kuroki, "Effects of AM-to-PM Conversion of TV Transmitters on Video Signals", IEEE Trans. on Broadcasting, VOL. BC-22 #3, Sept. 1976, pp 57-67.

[5] T. Rzeszewski, "A System Approach to Synchronous Detection", IEEE Trans. on Consumer Electronics, VOL. CE-22, #2, May 1976, pp 186-193.

[6] F. Hilbert and N. Parker, "Technical Considerations for Stereophonic Broadcasting on TV", IEEE Trans. on Broadcast & TV Rec., VOL. BTR-11, #2, July 1965, pp 82-89.

[7] H. O'Neill, "Incidental Phase Modulation and Sound Channel Noise in Intercarrier Systems", E&D Report 106/72, Engineering Division IBA (Great Britain).

[8] S. Deutsch, Theory and Design of TV Receivers. New York, McGraw Hill, 1951, pp 465-467.

[9] S. A. Roth and C. W. Rhodes, "A New Precision TV Measurement Demodulator", IEEE Trans. on Broadcasting, VOL. BC 23, #4, Dec. '77, pp 109-112.

[10] C. W. Rhodes, Testing and Using Synchronous Demodulators, Tektronix TV Products Application 28, 1979.

[11] C. Eilers and P. Fockens, "Television Multichannel Sound Broadcasting - A Proposal" IEEE Trans., Vol. CE-27 #3 August 1981, pp 398-409.

[12] P. Fockens and C. Eilers, "Intercarrier Buzz Phenomena, Analysis and Cures" IEEE Trans., Vol. CE-27 #3. Aug 1981, pp 381-397.

IPM DUE TO PARAMETRIC NONLINEARITY

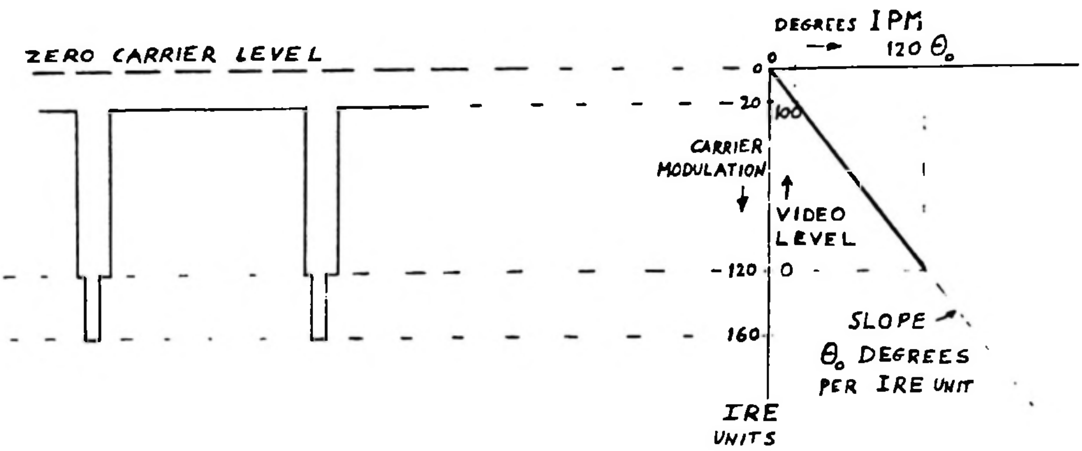


FIG. 1

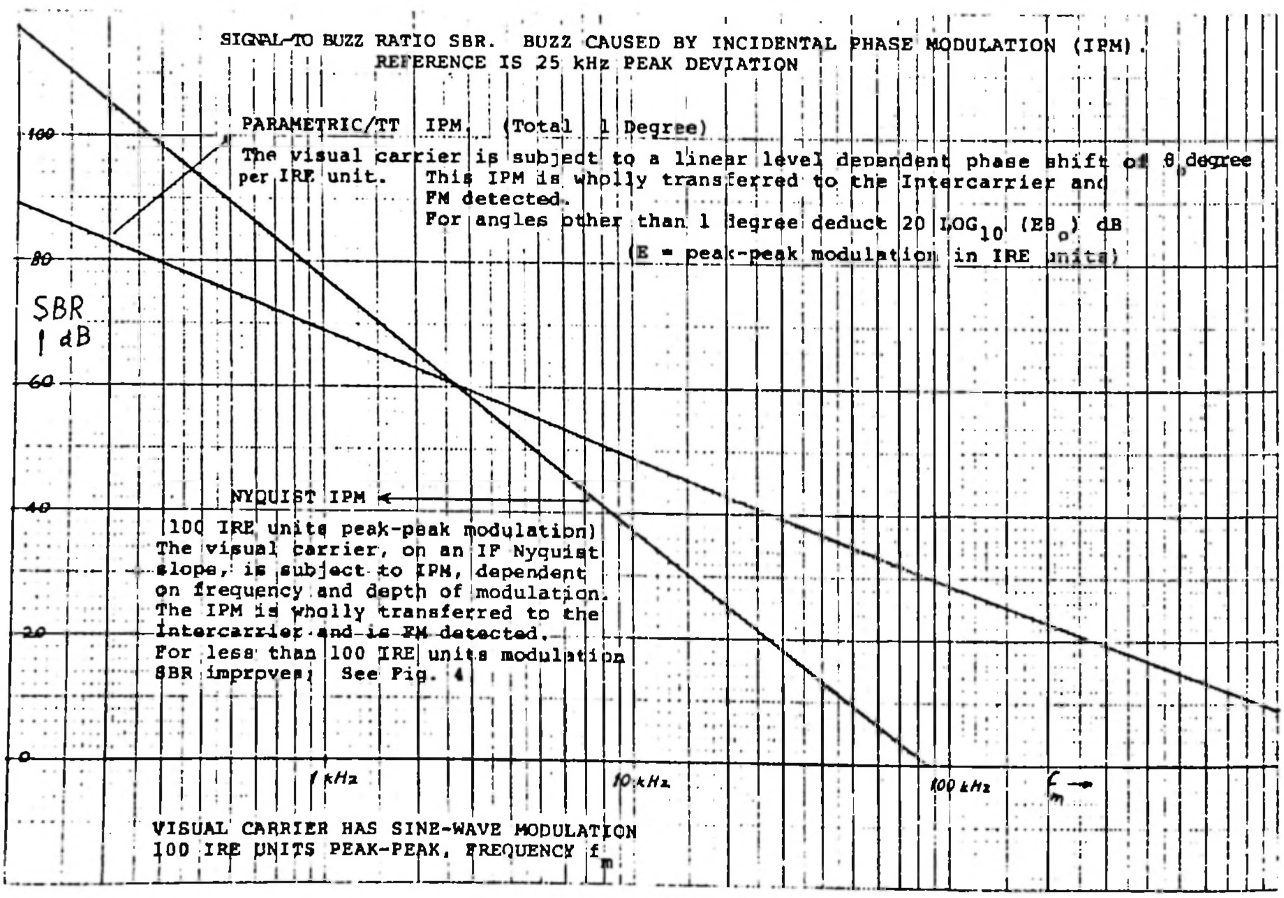
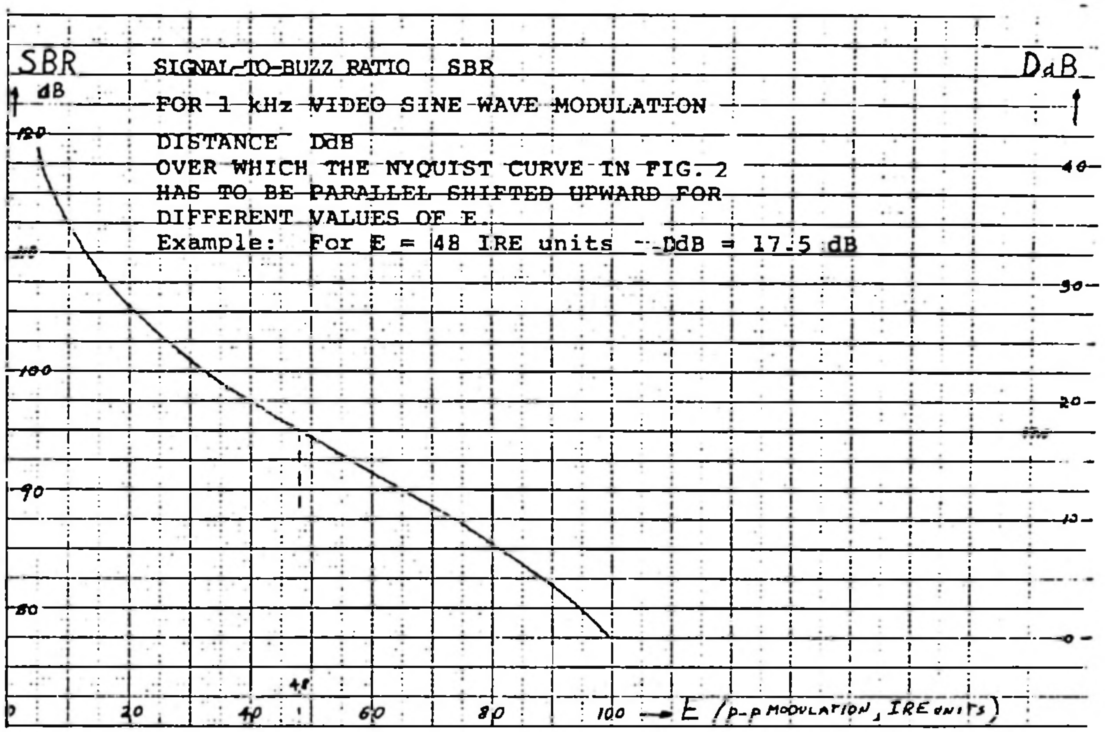
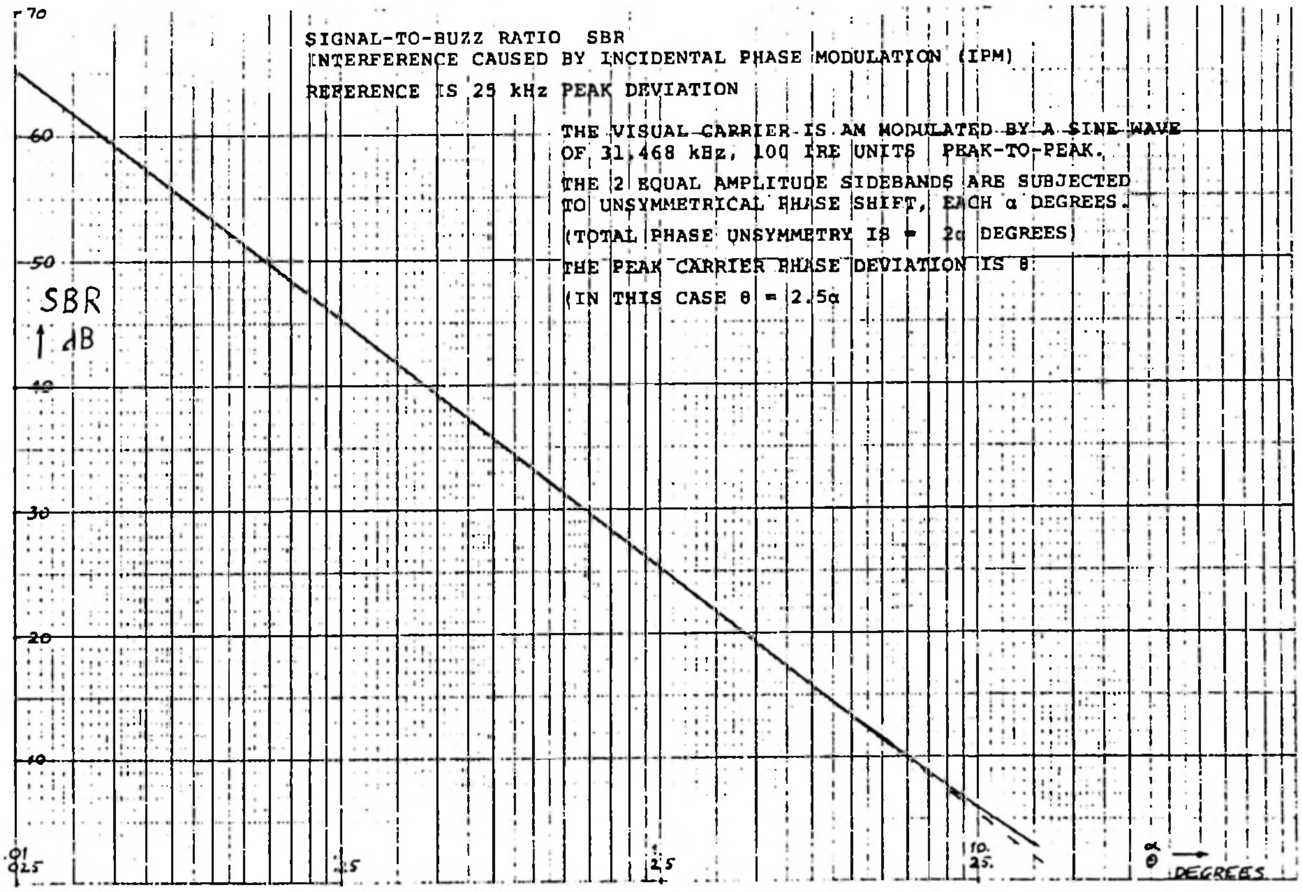


FIG. 2



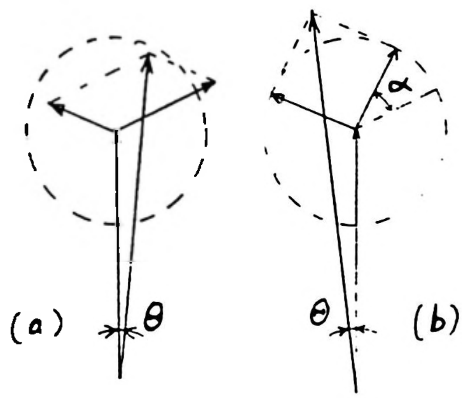
SBR
dB
SIGNAL-TO-BUZZ RATIO SBR
FOR 1 kHz VIDEO SINE WAVE MODULATION
DISTANCE DdB
OVER WHICH THE NYQUIST CURVE IN FIG. 2
HAS TO BE PARALLEL SHIFTED UPWARD FOR
DIFFERENT VALUES OF E.
Example: For E = 48 IRE units → DdB = 17.5 dB
DdB
120
100
90
80
40
30
20
10
0
0
20
40
48
60
80
100
→ E (p-p MODULATION, IRE units)

FIG. 4



SIGNAL-TO-BUZZ RATIO SBR
INTERFERENCE CAUSED BY INCIDENTAL PHASE MODULATION (IPM)
REFERENCE IS 25 kHz PEAK DEVIATION
THE VISUAL CARRIER IS AM MODULATED BY A SINE WAVE
OF 31.468 kHz, 100 IRE UNITS PEAK-TO-PEAK.
THE 2 EQUAL AMPLITUDE SIDEBANDS ARE SUBJECTED
TO UNSYMMETRICAL PHASE SHIFT, EACH α DEGREES.
(TOTAL PHASE UNSYMMETRY IS = 2α DEGREES)
THE PEAK CARRIER PHASE DEVIATION IS θ
(IN THIS CASE θ = 2.5α
SBR
dB
70
60
50
40
30
20
10
.01
.025
.1
.25
1
2.5
10.
25.
α
θ
→ DEGREES

FIG. 5



IPM DUE TO UNEQUAL SIDEBAND AMPLITUDE (a) OR PHASE (b)

FIGURE 3

SPLIT SOUND RECEIVER

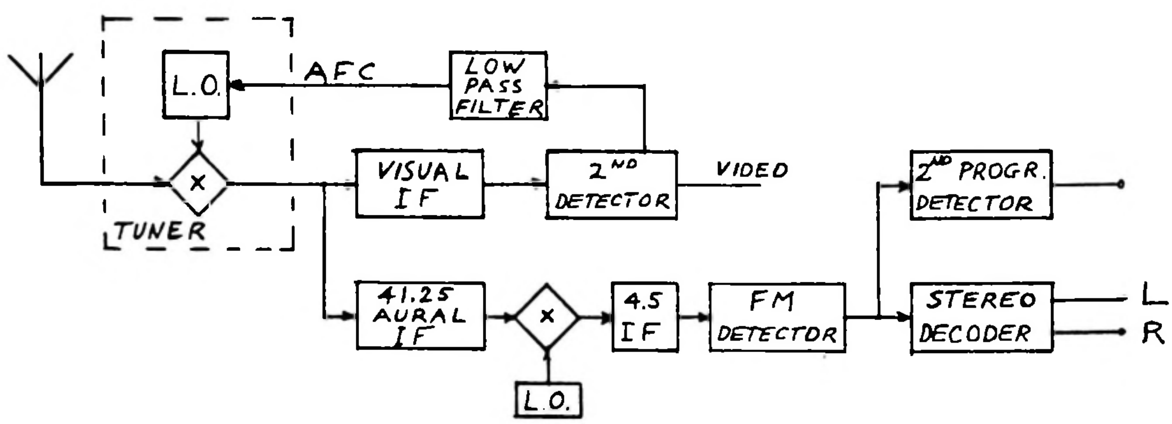


FIGURE 6

QUASI-SPLIT SOUND RECEIVER

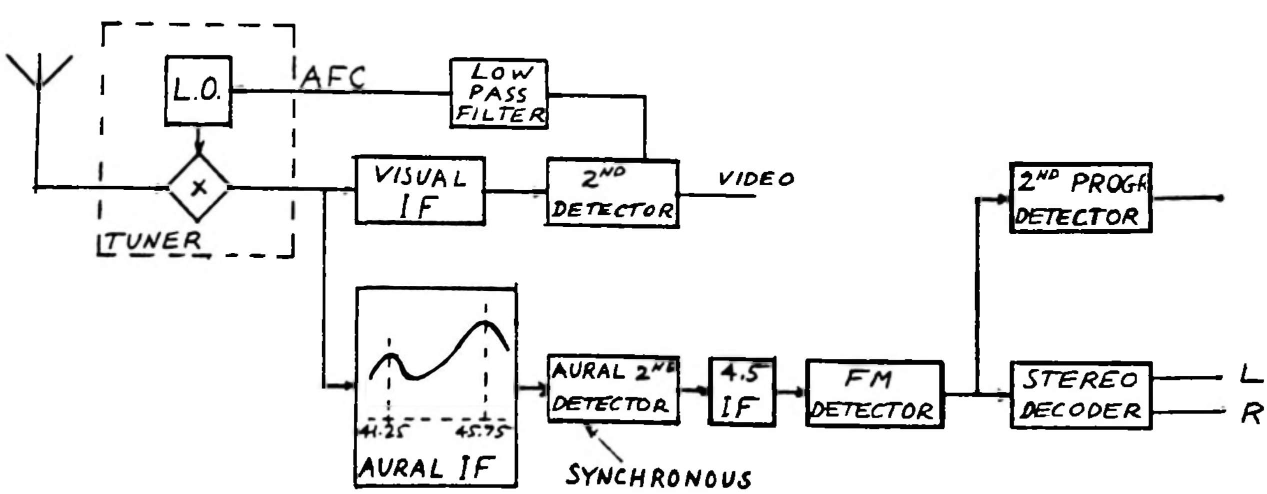


FIGURE 7

The Transmission Of Multichannel Television Sound

Through Cable Systems

Alex Best

Scientific-Atlanta

Atlanta, Georgia

1. Audio Signal/Noise Ratio[1]

In Volume I of the Multichannel Television Sound report, it was indicated[2] that for high quality stereo sound reception, the signal to noise ratio should be at least 60dB, preferably 70dB or better, for the principal community of viewers. For comparison, EIA RS-250B specifies for monaural sound that the minimum unweighted audio SNR for end-to-end television relay facilities be 56dB, including buzz.

In cable systems there are no boundary conditions between the principal and non-principal viewers therefore it must be considered imperative that we deliver at least a 60dB SNR to the TV receivers connected at the extremities of our systems. To determine the expected SNR (thermal noise only) in cable systems, calculations were performed by several members of the NCTA sub-committee. Although the results of these calculations differed by several dB, the conclusion to be drawn remains the same.

For the case of 36dB NCTA Video RFSNR[3], with the sound carrier 15dB below the video peak envelope power, we get the following unweighted audio SNR's.

[1]In order to simplify analysis and due to similarities between systems the Zenith was used for discussion and analysis except where clear differences are noted.

[2]Multichannel Television Sound: "The Basis for Selection of a Single Standard", Volume I, The National Association of Broadcasters, July 16, 1982, p.51.

[3]Minimum allowed CNR for Cable Systems. Part 76 of FCC Regulations.

	Monophonic	Stereo L or R	Separate Pgm
Separate Mixing	63.8dB	49.2	39.6
Intercarrier, Video at Blanking Level	63.6	49.0	39.4
Intercarrier, Video at White Level	59.0	44.4	34.8

Obviously we are in trouble if there are subscribers whose ears demand stereo but whose eyes will tolerate 36dB. In addition, the Separate Audio Program SNR will be approximately 10dB less than stereo signal. If we assume that a more typical situation is for the cable system to deliver a 43dB Video SNR, then we find ourselves dealing with a stereo SNR of approximately 55dB and a SAP SNR of approximately 45dB. These are certainly more tolerable conditions and could be improved even more if a successful companding system was found. For this reason the NCTA urges the EIA committee to continue its efforts to locate and approve such a system.

From the calculations it appears that threshold margins are adequate, both for new sets with sound IF bandwidths sufficient to support the broadband signal and for old sets. If cable systems get careless and allow the sound carrier to operate 20dB below the video carrier, this may not be the case. First to go would be the SAP in new sets which have that provision.

2. Visual/Sound Ratio

It is recognized that one way to improve the multichannel sound SNR is to increase the sound carrier level. Above threshold the audio SNR is increased one dB for each dB we raise the sound carrier level. In Volume I[4] of the EIA report, use of the highest feasible sound power is encouraged.

It must also be recognized by the EIA committee that increasing the sound carrier level in a cable system is an unacceptable recommendation, creating a multitude of problems, as explained in sections a. through d.

a. TV Receiver Adjacent Channel Rejection

All cable systems operate with TV channels adjacent to one

[4]Multichannel Television Sound Report, p.164.

another. In the early days of cable it was determined that most TV receivers lacked sufficient selectivity to provide for beat free pictures, unless the lower adjacent sound carrier was reduced in level below the visual/sound ratio being transmitted by broadcast stations. Although the NCTA committee recognized that receiver selectivity has been considerably improved over the years, the EIA committee must recognize that the proposed multichannel sound standards with their increased deviations (approximately 70KHz) creates sound carries whose energy occupies a much wider bandwidth than before (approximately 320KHz for 70KHz deviation as compared to 80KHz for 25KHz deviation). In a paper written in 1972 by Will Hand of Sylvania entitled "Television Receiver Requirements For CATV Systems", tests were performed on a number of color receivers which represented collectively over 50% of receiver designs in the industry at that time. The data shown below is the weighted results of these measurements. The data was weighted to reflect the proportion of the market held by the receiver manufacturers.

	Frequency(KHz)
Adjacent Sound null to 41dB point on low frequency side of IF trap	94
Adjacent Sound null to 41dB point on high frequency side of IF trap	115
The 41dB rejection bandwidth of this trap is 94 + 115 = 209KHz	

One can only conjecture what the rejection of this trap would be to a sound carrier adhering to the proposed multichannel sound standards; however, it goes without question that raising the sound level would only aggravate what already may be an unacceptable operating condition.

Set-Top converters with adjacent channel traps would help, although only a few presently in the field have this feature. Separate traps connected to the output of converters and tuned to the lower adjacent sound would help; however, they suffer from the following problems:

1. AFC and fine tuning errors would diminish effectiveness.

2. They would introduce additional group delay errors.

3. The overall cost would be substantial.

4. Subscribers might confuse them with pay TV traps and remove them.

b. Headend Equipment (Signal Processors, Strip Amps, and Modulators)

The majority of these devices operate at an output level between +50dBmV to +60dBmV (+60dBmV = 1 volt rms @ 75 ohms) for the video carrier with the sound carrier typically 15dB below this level. Most manufacturers also quote a specification which states that when the unit is operated at maximum output (this is not uncommon), all spurious outputs will be at least 60dB below the desired video carrier. One particularly troublesome spurious signal is created by third order intermodulation between the video and sound carrier. One component of this distortion falls 1.5MHz above the video carrier of the lower adjacent channel. Any attempt to raise the sound carrier level causes a one dB rise in this undesired signal for each dB the sound level is increased. This is a particularly sensitive area of the video spectrum and any interfering carrier must be 55dB to 60dB below the desired carrier not to be perceptible. This problem is only aggravated when scrambling systems are employed which amplitude modulate the sound carrier with descrambling timing information. The NCTA recognizes that this potential problem can be solved with highly selective bandpass filters on the ouptut of these devices but not without an increase in the envelope delay distortion inherent in these filters.

c. AML Equipment

AML equipment is used to transmit multiple channels of television from one area of a community to another. Increases in aural carrier levels would cause corresponding increases in the lower 4.5MHz products, as described in b. above. For systems operating at full rated power, it would likely be necessary to reduce output power thereby reducing fade margin. This equipment is also plagued with scrambling systems which cause a 6dB increase in the aural level during transmission of descrambling timing pulses. This practice creates a marginal condition to exist with today's practices. Increasing levels for Multichannel Television Sound would be unacceptable.

d. Distribution Equipment

Measurements were conducted to determine the perceptibility of sound carrier beats when the sound levels were increased above 15dB below the video carriers. From these measurements, perceptibility of sound-carrier beats occurs at a system level 3 to 3½dB above that level which produces barely perceptible video carrier beats (CTB-composite triple beat distortion) when not phase locked. In the same test the system level could be elevated approximately 5dB when phase locked before background images (modulation cross-over) could be seen. These tests were performed with 54 channel HRC loading. The conclusion to be drawn from these tests indicate if aural levels must be increased to accommodate Multichannel Sound, then the advantage gained from phaselocking is lost.

A second problem surfaced if aural levels are increased. Many cable operators find it possible to use certain channels in the aeronautical radio service bands by lowering aural carrier levels to +28.75dBmV maximum. In fact, these channels can sometimes only be used this way due to conflicts with both the visual and aural carriers. If aural levels are raised, this practice would be eliminated resulting in the loss of these channels for stereo service.

3. Increased Deviations

In order to achieve the highest possible stereo SNR, plus provide for auxiliary services, all proponents of the multichannel sound systems intend to increase the peak deviation to approximately 70KHz. By doing so, technical problems are created in both cable headends plus set-top converters, not to mention the selectivity problems of TV receivers. The following section discusses the nature of these problems in detail.

a. TV Receiver Selectivity (See section 2-a.)

b. Headend Equipment

1. Signal Processors

Two types are currently in use. One type uses a split sound system where the aural carrier is trapped and processed separately. The second type processes the video and sound combined and uses adjustable traps to reduce the sound carrier level. Both systems will suffer when deviations are increased from 25KHz to 70KHz.

For split sound units, the sound notch must not introduce amplitude or group delay errors in chroma information while attenuating the sound carrier and its sidebands at least 40dB. This has been achieved for 25KHz deviation with a Carson bandwidth of 80KHz. When the new Carson bandwidth for stereo (approximately 350KHz) is measured, attenuation of the upper sideband is on the order of 10dB. With incomplete trapping, a portion of the aural carrier passes through the visual processing circuitry. When this signal recombines with the processed aural signal, impairment of the received aural signal will result. This impairment could result in a substantial reduction in the amount of separation between the left and right channels. More testing needs to be conducted to confirm or deny this concern.

One processor measured had 2% AM on the FM sound carrier with 25KHz deviation and 16% AM with 70KHz deviation. This would clearly cause a problem when using analog descrambling information on the aural carrier.

In addition to the notch problem, the 3dB bandwidth of the sound path was measured on one type of processor to be approximately 350KHz. No attempt is made to control delay characteristics at the band edge. One member of the committee described the use of unequalized filters in home stereo receivers. This practice results in stereo separation of 40dB, a number considered adequate. General practice has been to control amplitude characteristics of filters, but no special care is used in controlling delay characteristics. This practice has not been reported to have caused any problems of which the committee is aware.

In general, the NCTA sub-committee agreed that IF circuitry in all existing processors would have to be redesigned for successful MCS operation.

2. Modulators

If external MCS signals are generated and pre-emphasis circuits are removed from modern modulators, it may be possible to use them for MCS transmission without difficulty. Older designs may have problems caused by transformers coupling and uncontrolled amplitude and phase characteristics in filters.

3. Demodulators

Volume I discusses the impact MCS will have on TV receivers; since most demodulators use similar circuitry, they will experience many of the same problems. This includes insufficient sound IF bandwidth, problems with buzz caused by intercarrier sound processing and inadequate baseband response to pass MCS subcarrier components.

c. Set-Top Converters(Scrambling/Descrambling)

For all the possible problems this one has the potential for the greatest impact on the CATV industry. Devices with the greatest proliferation use either pulse or sinewave sync suppression. Problems with both systems can arise when FM to AM conversion occurs in headend processors, modulators, and set-top converter bandpass filters. These products manifest themselves in different ways. In the case of pulse systems, the AM pulse on the sound carrier, and any spurious AM products, fire a trigger circuit, usually a monostable multivibrator. Any stray AM products or noise can cause descrambling pulse jitter due to slicer uncertainity.

An attempt was also made to simulate stereo susceptibility to stray tagging and descrambling pulses. Measurement data seems to indicate compatibility between MCS and pulse scrambling systems.

Of special concern in this regard was the presence of any stray information at 15750 Hz. The committee concluded from these measurements that the 15750 Hz components which we generated would be small compared to the levels normally encountered as a result of the hostile environment of a TV receiver. Several manufacturers use additional pulses on the sound channel for tagging; the effect these pulses might have on pilots was not investigated.

The sinewave sync suppression system may be most susceptible to FM to AM conversion products. These systems will undoubtedly suffer from the almost certain increase in AM products on the aural carrier. Additional testing needs to be performed in this area.

Several manufacturers have baseband systems in the field. Since these units have demodulators, the performance and problems will be similar to those experienced with TV sets. Special MCS units will likely be similar to the tuner, IF amp and detector circuitry developed for TV receivers intended for MCS operation. This suggests units already in the field can continue to be used for subscribers with monophonic TV receivers, while units of a new, compatible design would be required for subscribers wishing to avail themselves of the opportunity to have multichannel sound. The one unknown, to be investigated, was discussed earlier in section 2-a of this report. The adjacent channel sound traps in these products could be expected to behave like those in TV sets. One manufacturer uses crystal stabilized local oscillators with AFC and SAW IF filters. When compared with older tube type television sets with unstable IF circuitry, these units may offer satisfactory performance. This must be verified before we relax our concern.

4. Phase Noise Considerations

In Volume I of the Multichannel Sound Report, mention was made that a return to split-sound TV receiver design could eliminate sound buzz due to incidental phase modulation of the sound carrier generated by mixing with the video carrier. This would certainly be the case; however, the EIA committee and the TV receiver manufacturers must also recognize that there are additional problems created by this technique which most likely would be worsened by equipment presently used in cable systems. Indicated below are several areas which need additional investigation before split-sound receiver design is seriously considered.

a. AML Equipment

On page 161 of Volume I of the MCS report, the EIA committe discusses a problem with translators in Japan. "During the rebroadcast tests of MCS signals, a large increase in buzz interference within the subchannel was observed. The interference was especially noticeable in split-carrier reception. This was due to amplitude-phase nonlinearity of traveling-wave-tube (TWT) power amplifiers used in old translators and also due to cascaded use of such translators." There is reason to believe a similar problem may exist with AML equipment.

Although most operators use phase locked receivers for AML transmission, some operators occasionally use receivers with free running local oscillators. In such cases, the levels of IPM increase substantially. One cable operator measured substantial increases in noise levels using these receivers with synchronous demodulators. MCS TV receivers using split sound systems would experience impaired noise performance as a result of this phenomonon.

b. Headend Equipment

Most signal processors and modulators use oscillator designs which are crystal controlled; however, there are presently on the market several multichannel units (typically used for standby purposes) which use oscillators that are either frequency synthesized or under AFC control. These devices will almost certainly increase the amount of incidental phase modulation present on the sound carrier.

c. Set-Top Converters

Converters which use synthesized tuning systems introduce substantial levels of incidental phase modulation (ICPM) on the TV signal. Older varactor tuned products also introduce ICPM, but to a lessor degree. TV receivers using intercarrier sound detection are not affected by high levels of ICPM when both the visual and aural carriers are subjected to the introduction of ICPM; however, in TV receivers using split sound IF and detection systems ICPM products introduced by CATV converters can contribute to deteriorated noise performance. It is not uncommon for varactor converters to introduce frequency modulation on the TV signal in excess of 10KHz. This suggests a serious problem for the cable industry if split sound TV receivers are introduced in the marketplace.

This paper has described several areas in cable systems where the carriage of multichannel sound might create problems. As of the writing of this paper the NCTA Engineering Committee has outlined a test plan to investigate these areas. These tests will be performed by representatives of the cable industry working in conjunction with the EIA test engineers. This testing will begin in February and last about 6 months. The results of these tests will be a comprehensive report including both subjective and objective evaluations of each area of concern.

VISUAL AND AURAL EXCITER DESIGN

FOR MULTI-CHANNEL SOUND

ROBERT M. UNETICH

INFORMATION TRANSMISSION SYSTEMS CORP.

CANONSBURG, PENNSYLVANIA

The transmission of aural subcarriers over existing and new television transmitters requires certain transmission characteristics of both the visual and aural portions of the transmitter to be improved or changed.

The transmission of stereo and a second audio program requires multiple high quality subcarrier systems, with excellent signal-to-noise ratio, low cross-modulation, low distortion, high deviation capability and wide audio bandwidth. Since these characteristics are helpful to the successful transmission of nearly any subcarrier system, the transmitter design needs to consider these and other parameters if general subcarrier capability is to be provided.

Since high power amplifiers often introduce ICPM distortion, incorporating this subcarrier capability into existing transmitters requires knowledge of the characteristics of the higher power stages in an RF system and then the selection and installation of a suitable "exciter". This exciter must be capable of providing the basic modulation capabilities for the subcarriers to be employed, filtering to provide sufficient separation of video and audio components and the correct precorrection characteristics to compensate for amplifier distortions and drifts. Additionally, as an integral part of the new RF system, it must often interface with other related subsystems such as remote control and mod-anode pulsers.

The present EIA committee work going on in selecting a multi-channel sound system for high power broadcast TV use has generated an array of general characteristics that an RF system should have to allow its use with the multi-channel sound systems. From this general requirement of the full system and from experience with VHF and UHF high power systems, a specification for exciter performance can be developed. While final determination of RF system requirements has not been published by the committee studying multi-channel sound, general requirements have been widely discussed and the author's interpretation of these requirements is listed in Table 1.

MCS
RF SYSTEM REQUIREMENTS

AURAL

Bandwidth	50 Hz to 110 kHz
Response	$\pm$0.5 dB
Deviation Capability	$\pm$75 kHz
Harmonic Distortion	0.5%
Audio Phase Shift (from linear)	50 Hz to 60 kHz: 3°
FM Noise (relative to $\pm$25 kHz deviation)	50 dB

VISUAL

ICPM	$\pm 2^{\circ}$
Video Response	$\geq$13 dB attenuation above 4.4 MHz

This list is not intended to delineate all system specifications, only those significantly relating to MCS service.

Table 1

These characteristics are achievable with present circuit technology but are not present as standard characteristics in most of the transmitters now in use. These RF system requirements can be met at both VHF and UHF with pre-correction and consideration of system drifts. Correction of distortions at VHF is often different than at UHF where most system non-linearities are present in the klystron amplifiers. Additionally, VHF exciters vary greatly in power output level requirements ranging from tenths of a watt to hundreds of watts. A UHF exciter is generally operated from 0.5 watts to 10 watts in output power and often a mod-anode pulser is driven by or connected to the exciter. Interfacing with a pulser includes providing the proper timing. Ideally, the long delay typically present in SAW VSB IF filters should be managed without complicating the pulser interface. This can be accomplished by developing the sync pulser drive signal from a point in the exciter after the SAW filter.

Klystron amplifiers introduce the largest amounts of ICPM, ranging from several degrees from white to sync tip, to as much as 30° additional phase shift when mod-anode pulsing is employed.

From the above, an exciter specification can be developed that will satisfy the system requirements. Of course, the practical electrical interface characteristics must be compatible with in-place transmitters, pulsers and remote control systems.

Combining these exciter requirements with other generally desirable exciter features allows a general exciter specification to be developed and for a block diagram to be developed for such a system. A system block diagram which satisfies these requirements for UHF is shown below in Figure 1.

As can be seen in the block diagram, the aural exciter circuit includes the following functions:

* Balanced baseband audio input with pre-emphasis
* Special 20 kHz to 150 kHz subcarrier input (75 ohms unbalanced)
* Composite audio input jack with no pre-emphasis (75 ohms unbalanced)
* 41.25 MHz voltage controlled oscillator (FM modulator)
* Phase-locked loop frequency control, locking the VCO center frequency to the 45.75 MHz crystal oscillator.
* Aural level control
* Aural IF interface jacks for STV scrambling

The aural IF signal is routed directly to the Upconverter tray.

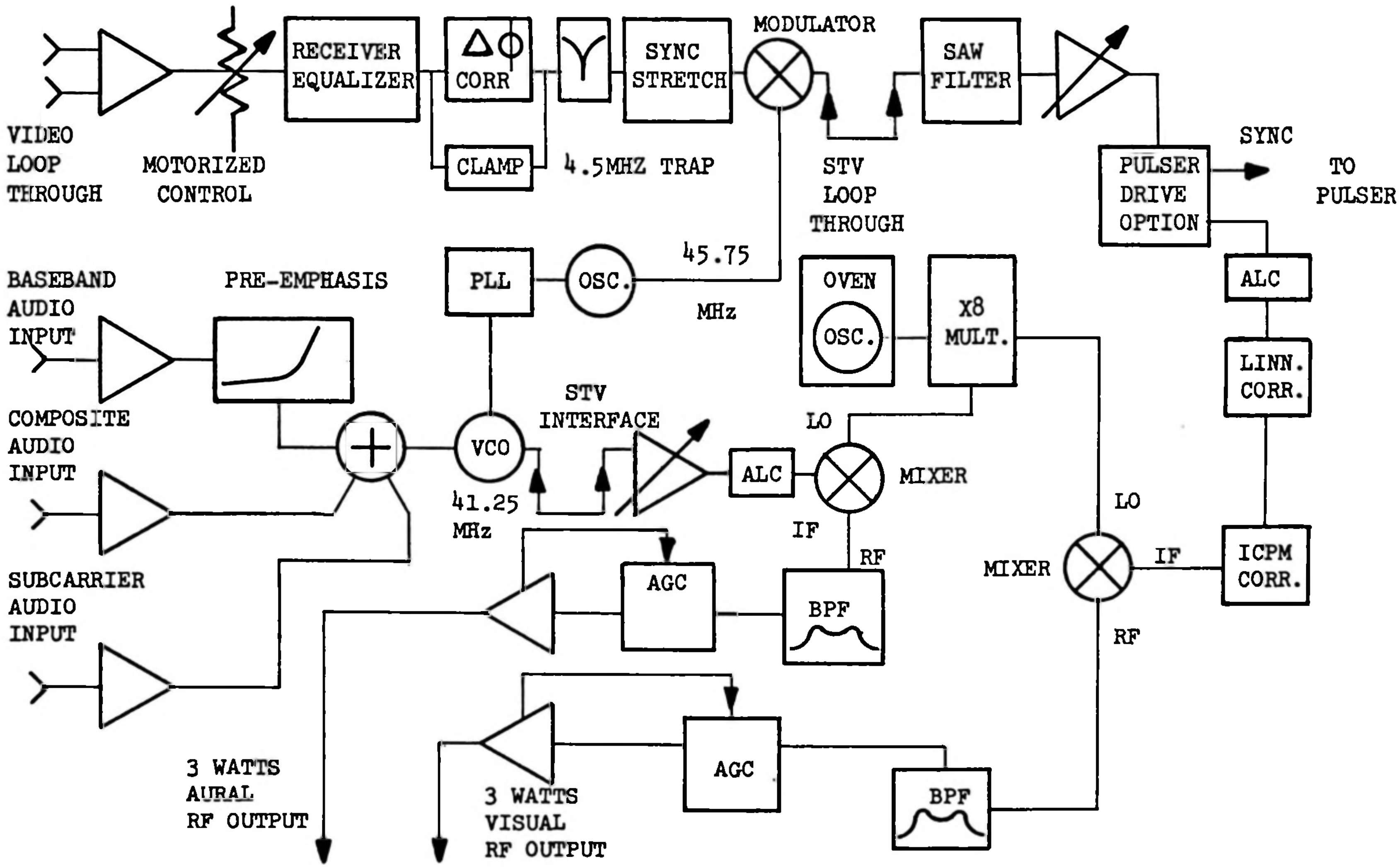
The video signal is processed by the following functional circuits in the modulator tray:

* High impedance loop through input
* Video level control
* Back porch clamp
* Differential phase corrector
* 4.5 MHz video trap
* Buffered, bypassable, receiver equalizer
* Motorized video level control
* Video sync stretcher
* Balanced modulator with ICPM nulling
* IF loop through for STV scrambling
* Bypassable SAW filter

After leaving the Modulator, the visual signal is routed to the optional Pulser Interface tray which is only used in systems that employ mod-anode pulsing. In this tray, the IF signal is synchronously demodulated and sync is separated from active video. A delay network in the IF path is provided to insure the proper timing of pulser operation. Pulser drive circuitry is provided and a switchable delay network is included in the pulser drive path to fine-adjust the timing.

Since the pulser drive is developed after the SAW filter, the time delay of the SAW device is not a factor in system operation. This eliminates the effect of temperature drift and allows bypassing of the SAW device with the pulser operating for system testing.

The output of this tray or the modulator tray, depending on the use of pulsing, is sent to the UHF Upconverter Tray. As can be seen on the block diagram, this tray includes the following functions:



SIMPLIFIED BLOCK DIAGRAM
ITS
TELEVISION EXCITER
FIGURE 1

* IF Automatic Level Control with motorized interface
* Multiple sections of IF amplitude linearity correction
* IF incidental phase correction with synchronous demodulation of an IF sample and video shaping.
* Oven controlled crystal oscillator
* Broadband multiplier
* Appropriate amplifiers, mixers and filters for upconversion of the IF signals.

The outputs of this tray are at about the 100 milliwatt level. These signals are then amplified to about the 3 watt level in separate amplifier trays. Each of these trays include power supplies, variable AGC circuits and logic control interface circuits. The amplifiers are Class A designs for high stability and linearity. Additional higher power klystron driver trays are also available up to 10 watts peak sync.

These five trays comprise a complete UHF exciter capable of driving most existing transmitters and interfacing well with existing mod-anode pulsers. In systems without pulsers, the complete exciter is comprised of four trays.

Mechanically, 19 inch rack mount trays are employed allowing easy installation in available standard cabinet space. All of the trays are 3½ inches high, resulting in a full system requiring only 17.5 inches of panel height. Cable retractors and slides are a necessity for rack mounting and are included in the product design.

All circuit boards are accessible through tray top access. LED's are provided in numerous circuit locations as status indicators. Remote control interface connections are made at a "D" connector in the rear of the exciter, attached to the exciter cable harness.

This design will provide the ability to operate a UHF transmitter system within the limits previously listed. As subcarrier systems for multi-channel sound or other "information" channels are perfected, this exciter design will allow the early implementation of these services.

AN UPDATE ON THE TECHNOLOGY

OF LOUD COMMERCIAL CONTROL

RALPH A. HALLER

FEDERAL COMMUNICATIONS COMMISSION

COLUMBIA, MARYLAND

AN UPDATE ON THE
TECHNOLOGY OF
LOUD COMMERCIAL CONTROL

EXECUTIVE SUMMARY

The Federal Communications Commission (FCC) annually receives numerous complaints of loud commercials being broadcast on radio and television. Over the years the Commission has issued various policy statements on steps that broadcasters can take to help reduce the loudness level of commercials (1, 2, 3). Unfortunately, because individual listeners vary in their judgments of the loudness(4) and since no adequate electronic metering devices have been available, precise control or regulation has been impractical.

The CBS Technology Center in Stamford, Connecticut, has been working for many years to develop a method to 1) measure loudness, and 2) automatically control loudness. The CBS work has led to two FCC evaluations of devices developed by CBS. In 1977, the Field Operations Bureau conducted a brief study of an early CBS loudness monitor. The second study, the subject of this paper, reviewed the operation of a "smarter" loudness indicator and an automatic loudness controller developed by CBS.

The new loudness indicator showed a high correlation with observers' responses to stimuli such as speech(5) and commercials. More variables were present in the loudness controller tests than in the loudness indicator tests making the results of that test less exact; however, the controller appeared to help reduce the perception of excessive loudness. Widespread adoption of such loudness control techniques could lead to fewer complaints of loud commercials.

BACKGROUND

Over the years, broadcasters and the Federal Communications Commission have received numerous complaints of loud commercials being broadcast on radio and television. This is evidenced, in part, by over 16 volumes of responses to the current loudness inquiry (3). Little was known about either what makes commercials loud or how to measure and control loudness levels. It became obvious in the early days of studying the problem that existing measurement techniques, such as the Volume Unit indicator (VU meter) or the modulation monitor, do not adequately measure perceived loudness.

As far back as the mid-1950's, attempts were made to measure and control loudness. General Electric's early attempt to develop such a controller resulted in a device called the "Unilevel" amplifier. The amplifier maintained the average audio voltage of the program signal at an established level. Similar devices have gained almost universal acceptance today and are installed in almost all broadcasting facilities at this time. Additionally, most stations use limiting amplifiers to prevent overmodulation. Unfortunately, these types of amplifiers do not respond to the apparent loudness of the audio signals, they respond only to voltage levels. Although other attempts to measure and control loudness have been tried over the years, the automatic control of loudness has remained the primary goal of such work.

The question of loudness level control became so prominent in the 1960's that the FCC issued a Notice of Inquiry(1) and a Public Notice(2) statement to broadcasters that indicated several steps to be taken to limit the loudness level of commercials. For example, the notice advised stations to avoid "compressing" audio by more than 6 decibels (dB). The FCC's Public Notice(2) did not lead to effective control of apparent loudness levels for at least two primary reasons: 1) program material constantly changes and cannot be adequately handled by manual level controls, and 2) loudness is highly variable and complex (4, 6).

During this same time period, the CBS Technology Center, Stamford, Connecticut, was conducting research aimed at learning specifically how to measure loudness levels, and how to control them.(7, 8, 9) The first result of their effort was a loudness level monitor (LLM) with an analog meter display similar to a traditional VU meter. This monitor represented an early attempt to develop a device which responded to sound in the manner of a typical "standard observer." The original CBS device did not achieve widespread acceptance by broadcasters, although a few units are in service to this day.

In 1975, the FCC, based on complaints received, again felt that loud commercials were becoming a problem. This prompted another Public Notice(11) to provide known steps to help broadcasters control loudness levels and put broadcasters on notice that loud commercials should be controlled in the public interest. In a later response to this Notice, CBS made one of the early design loudness level monitors(10) available to the FCC for testing.

William H. Hassinger, of the FCC's Field Operations Bureau, tested the LLM against responses from a small panel of observers using a video tape of actual television programming as source material. The results of the test were extremely encouraging in that a high correlation between the meter and the observers was found (12). Mr. Hassinger's work was published by the FCC in 1978 along with another Notice of Inquiry concerning the loudness question (3).

With encouragement from the FCC and its own interest in loudness, CBS instituted a new research and development program in 1978 to improve the loudness algorithms used in the loudness meter. It was recognized that any electronic device would never achieve 100% accuracy; however, new data, especially on frequency and amplitude sensitivity of the hearing mechanism and temporal integration times, led to the development of improved algorithms for use in the design of a new device---the Loudness Indicator (13).

CBS continued its work by merging the loudness algorithm of the indicator into an automatic loudness controller. A prototype controller was supplied to the FCC in 1981. Evaluation of the loudness controller was undertaken at the FCC laboratory during 1981 and 1982 by the Experimental Engineering Branch.

DESCRIPTION OF NEW LOUDNESS INDICATOR
(Supplied by CBS)

Loudness Summation

A distinguishing feature of the original CBS Loudness Level Monitor was its ability to perform loudness level additions on wide spectrum signals. To a first approximation this was in accordance with human neural response to vibrations of the basilar and tectorial membranes in the inner ear. The modeling of the basilar membrane as a filter with maximum transmission of different frequencies roughly localized at various points along its length has been investigated by many researchers. This has resulted in a refined understanding of so-called "critical bands." By accepted definition, a critical band is that maximum bandwidth below which loudness does not diminish as the bandwidth is decreased (4, 14).

An ideal loudness measuring device should sample program energy in each critical band prior to further processing of and/or display of the data. However, this would require a total of twenty-four filters and would result in an excessive component cost. In the current development program, the critical band concept was used, but with each three adjacent critical bands combined into each of eight broader filters rather than the twenty-four. Figure 1 shows the location of the theoretical critical bands and their combination into the broader band filters for the new Loudness Indicator.

Sensitivity versus Frequency

Frequency sensitivity and loudness summation represent the two functions most critical for the accurate performance of a Loudness Indicator. Normal hearing response at conversational levels has maximum sensitivity in the region from 2 to 8 kHz, and a progressively poorer sensitivity towards the lower and higher frequencies (4). In Figure 2, the sensitivity of the Loudness Indicator is shown by the upper curve. Below this curve, the response of the three-critical-band filter sections are displayed according to their relative sensitivity. The design of the Indicator is such that an arithmetic addition of the rectified outputs of the filters driven with wide-band noise signals results in the upper frequency response curve.

Temporal Integration

A loudness measuring device must be able to account for the acoustical-mechanical-neural response times of the hearing mechanism. It has been determined that the output of each filter can be appropriately detected with time constants of 20 msec attack and 200 msec decay. These filter outputs are subsequently summed and integrated with an overall attack time constant of 70 msec and a decay time constant of 350 msec. In the circuit employed, these two constants match the 200 msec temporal integration time of loudness growth

(15).

Circuit Design

Figure 3 shows a block diagram of the new Indicator. The front-end circuitry includes a bridging input transformer and a calibration/sensitivity control. Spectral division by eight band-pass filters, as previously described in Figure 2, initiate the loudness measuring algorithm. Full wave detection of the output of each filter is accomplished in conjunction with compensation circuitry for linear rectification over the entire dynamic range of the instrument. The loudness addition function is provided by feeding the individual rectified outputs to a simple summing circuit. This circuit is followed by an integration network and a front panel display.

Display

Various display techniques were considered for use with the new Loudness Indicator. Major factors considered included the need for accuracy of calibration, good visibility, and freedom from response time constraints which would be at variance with intended duplication of pyschoacoustic responses. This latter consideration appeared to rule out most electro-mechanical devices of the moving-pointer variety as being too sluggish for this application. The final choice utilized light emitting diodes (LED). The display utilizes fifty-one LED's extending to a length of five inches horizontally as shown in Figure 4. Forty yellow LED's give 1/2 dB resolution from -20 to 0 dB. Ten red LED's give similar resolution from 0 to +5 dB. A single green LED mounted below -- 20 dB acts as a pilot to indicate that the Indicator is energized. A linear dB format was chosen for the primary calibration. A secondary calibration utilizes traditional pyschoacoustic format, where "one-half loudness" or "twice loudness" is equivalent to 10 dB (6).

DESCRIPTION OF LOUDNESS CONTROLLER
(Supplied by CBS)

The CBS Loudness Controller performs two functions--loudness control and level control. It increases or decreases the level of incoming program material to maintain uniform average loudness while simultaneously holding the output level below a preset limit to assure proper modulation levels or to prevent excessive reaction by peak limiters which may follow the Loudness Controller.

The Controller comprises four major circuits: a) the audio signal path, b) a Gain Hold circuit, which inhibits gain changes when the input falls below a preset level, c) the loudness control circuitry and d) the level control circuitry. Also included are a combinational logic circuit and a front-panel display which indicates relative gain applied by the unit. A block diagram is shown in Figure 5.

The heart of the loudness control section is a circuit which applies a loudness algorithm to the input signal. This circuit is similar to that used in the CBS Loudness Indicator. A block diagram of the circuit is shown in Figure 6. This circuit takes the following three factors into account: 1) frequency content, 2) signal bandwidth, and 3) signal dynamics.

Due to the conflict between the broadcaster's desire to maintain as high a power output as possible and FCC rules regarding modulation levels, the Loudness Controller has been designed to limit the maximum amount of gain which can be contributed by the loudness section. This limit is a function of the loudness-to-level ratio of a signal. For example, if a low-loudness, high-level signal were presented to a loudness controller which had no form of level control, the controller would increase gain as much as necessary to bring the signal up to a uniform loudness level. The resultant increase in level could over-drive the station's peak limiter and, under extreme conditions, cause possible overmodulation. But the addition of a level-controlling circuit makes it possible to restrict the maximum output level to some upper limit. This limit is set when the OUTPUT LEVEL control is properly adjusted.

Under normal conditions, the Loudness Controller will cause a slight loss in modulation level during loud program segments. This is in agreement with FCC regulations which require that modulation levels be "usually not less than 85% on peaks of frequent recurrence, but where necessary to avoid objectionable loudness, modulation may be reduced to whatever level is necessary." This acknowledges the trade-off between loudness and level--it is impossible to reduce loudness without some decrease in modulation level. However, it is possible to minimize the decrease in modulation level at the sacrifice of some loudness uniformity. This non-uniformity will only occur on material which was initially of low-loudness and high-level since the Loudness Controller's level circuitry is limiting the allowable gain. This does not interfere with the Controller's ability to reduce the level of loud material as much as necessary.

TEST PROGRAM

Two testing programs were developed. In both programs, subjects were asked to watch a 20 minute video tape that had 15 commercials intermixed with news programs. The subjects watched and listened to a television receiver in a room with a home-like environment. Each subject was seated at a distance of five times the picture height. In the first test, the subjects made choices of how loud each commercial was in relation to the program level. Five level designations (soft, slightly softer, program, slightly louder and loud) were suggested. Votes were registered by the subjects by depressing one of five switches in a minibox. The responses of the subjects were recorded on a multi-channel strip chart recorder. Outputs of the loudness indicator monitor were also recorded. Each subject watched the video tape twice, once with the controller out of the circuit and once with the controller in the circuit. This allowed a comparison of the subjects' responses with the controller in operation versus the controller out of operation. The second test was basically the same in terms of setup except that instead of having a five category choice of loudness, the subjects operated a volume control to attempt to make all programming and commercials the same loudness level. The relative level of the volume control was recorded on the strip chart recorder. Block diagrams of the test setups are shown in Appendices 1 and 2. Appendices 3 and 4 show the instructions given to each participant.

The strip charts from the first tests were analyzed for differences in subject responses with the controller in or out. The second test strip chart tapes were analyzed for the dynamic range of subject reponses with the controller in or out, the premise being that if the controller was effective, the subjects would move the volume control less. Sample strip chart recordings are shown in Appendix 5. There were nine participants in the Type I tests and five participants in the Type II tests.

TECHNICAL DETAILS OF TESTS

For both tests (refer to Appendix 1) the video output of a video cassette recorder was used to modulate a television signal generator operating on channel 3. The audio output of the video cassette recorder was routed through the CBS loudness controller and then to the television signal generator. A switch on the controller allowed having the automatic action of the controller action on or off. Loudness indicators monitored the output audio line from the controller and supplied analog "loudness" voltages to a two-channel strip chart recorder. An event marker channel was also available on the strip chart to record responses in the Type I tests. The radio frequency output of the television signal generator was then routed to a television receiver in a room having a home "living room" atmosphere.

In both tests the subjects adjusted the volume control on the television at the beginning of the tape for a comfortable volume. For the duration of the tape, for the Type I tests, the subjects responded to commercials by depressing one of five buttons to record a vote of the relative loudness of the commercials. The switches were connected to a microprocessor that supplied an output to an event marker channel of the chart recorder. Choices were recorded as one to five tick marks on the strip chart. The following format was used:

Level	Ticks
Soft	5
Slightly Softer	4
Program	1
Slightly Louder	2
Loud	3

In the Type II tests, the five switches were removed and replaced with a volume control, external from the television. The control was a double-ganged potentiometer with one section acting as a direct current voltage divider for one channel of the strip chart recorder to provide information on the relative position of the potentiometer. The second section acted as a volume control on the audio line to the television signal generator. Initially the potentiometer was set to mid-range and the subjects adjusted the volume control on the television receiver for a comfortable level. All additional level changes were made with the external ganged potentiometer. By comparing the strip chart of each subject's changes in the potentiometer setting to a calibration strip chart showing potentiometer position versus decibel attenuation, each response could be directly converted into a decibel range in volume.

RESULTS

It was anticipated that the effects of the loudness controller could be studied by comparing reponses with the controller in operation to those without the controller.

In the Type I tests, the subjects' responses, as recorded by the event marker of the strip chart recorder, were compared to the recorded indication of the new Loudness Indicator and the earlier Loudness Level Monitor. This was accomplished with the controller in operation and out of operation. Then, the responses for the controller in versus the controller out could be compared to determine if fewer loud or sightly loud choices were made with the controller operational. There was a psuedo-random choice as to whether the controller was operational during the first viewing or during the second viewing. Participants were not told of the choice made. This allowed study of whether a learning curve was apparent during the second viewing. However, it appeared to make no difference whether the controller was in during the first or second viewing. Early trials of the Type I tests did indicate that both viewings of the tape should be close in time due to the importance of having the subjects be in the same psychological state for both viewings.

While the differences between controlled and non-controlled levels were fairly easy to see on the chart recordings, it was not so easily demonstrated pyschoacoustically. The typical response range did shift toward a higher designation (i.e. toward "louder") when the Controller was not operating. Also, a few listeners made unsolicited comments to the effect that the levels seemed more uniform in sessions in which the Controller was operating.

An examination of the differences in readings of the Loudness Indicator between controlled and uncontrolled audio showed that the loudness levels of 60% of the test commercials were unchanged when the Controller was operative, suggesting that the original levels were acceptable. The Indicator also showed that the level of 27% of the commercials were decreased and 13% were increased by the Controller. In comparison, the judgment of the listeners suggested that 50% were unchanged, 35% reduced and 15% increased.

TABLE 1

COMPARISON OF LOUDNESS INDICATOR RESPONSES TO SUBJECTS' RESPONSES

Commercial Loudness	Listener Vote	Loudness Indicator
Same	50%	60%
Level Up	15	13
Level Down	35	27

For the Type II tests, each listener was given an external volume control and was instructed to keep the loudness level even. Five subjects were tested. Although the range of volume control movement varied considerably between observers, the differences in volume control movement for each observer with the controller in or out was not significant. Since it appeared that no further information was being gained by this test procedure, the subjective tests were terminated.

Both Test Types I and II suggested that the Controller was acting properly, but the small gain adjustments needed with the test program material used did not permit adequate test sensitivity for listener judgments. Source material for such tests should have widely varying loudness segments. This would have had yielded more information on the effectiveness of the Controller.

This testing program also permitted examinations of the Loudness Indicator's responses by comparing the direction of (louder or softer) the listeners' responses as shown on the strip charts to the direction of the Indicator's responses also shown on the strip charts. This comparison of the direction of change was made for every commercial under each condition (Controller in or out) for the listeners and the Indicator. There was 80% agreement overall for all listeners in all conditions, including preliminary results from the design stages of the Type I tests.

CONCLUSIONS

Work accomplished by CBS has significantly advanced the state-of-the-art in loudness measurements and control. The high correlation between the responses of the participants in the testing program and the levels displayed on the Loudness Indicator show that the loudness algorithm closely approximates human perception of loudness. Because this algorithm appears to be valid, and the algorithm forms the basis of operation for the controller, it follows that the controller should be helpful in the reducing the number of loud commercials and the number of complaints.

Even if it could be shown that the controller was 100% effective some complaints would still be received. Individual hearing response, specific content of commercials, visual displays for television manner of presentation and other psychological variables alter one's perception of loudness.

The results reported herein are based on a limited number of samples and may not be definitive in all cases. Also, opinions expressed herein are those of the author and do not necessarily reflect views of policies of the Federal Communications Commission.

REFERENCES

1. Notice of Inquiry, Docket 14904, "Loud Commercial Inquiry," 1962.

2. FCC Public Notice, July 12, 1965, Statement of Policy Concerning Loud Commercials, FCC 65-618, 69685.

3. Notice of Inquiry, Docket BC 79-168, "Loud Commercial Inquiry,", FCC 79-412, 14205.

4. S.S. Stevens, Phychophysics, New York: John Wiley & Sons, 1975, page 74.

5. CCIR, International Radio Consultative Committee, Tape recording for the determination of the subjective loudness of a boradcasting program. (Question 4/X) Draft Report E.1.k(x) ITU Geneva, 1969.

6. L.E. Marks, Psychophysical Measurement Procedures, Tasks Scales. In (B. Wegener edition) Social Attitudes and Psychophysical Measurement. Hillsdale, Lawrence Erlbauem Associates, 1982, pages 58-60, 49, 60.

7. B.B. Bauer and E.L. Torick, "Researches in Loudness Measurement", IEEE Transactions on Broadcasting, Volume AU-15, Number 3, September 1966.

8. B.B. Bauer, E.L. Torick, and R.G. Allen, "The Measurement of Loudness Level", J. Acoustical Society of America, Vol. 50, No. 2, August 1971.

9. E.L. Torick, R.G. Allen, and B.B. Bauer, "Automatic Control of Loudness Level", IEEE Transactions on Broadcasting, Vol. BC-14, No. 14, December 1968.

10. B.B. Bauer, E.L. Torick, A.J. Rosenheck and R.G. Allen, "A Loudness Level Monitor for Broadcasting", IEEE Transactions on Audio and Electroacoustics, Vol. AU-15, No. 4, December 1967.

11. Public Notice, July 29, 1975, reissuance of 1 above.

12. W. Hassinger, "Evaluating Loud Commercials: An Experimental Approach", FCC/FOB 78/01, February 1978.

13. B.L. Jones and E.L. Torick, "A New Loudness Indicator for Use in Broadcasting", Journal of the SMPTE, Volume 90, Number 9, September 1981.

14. L.E. Marks, "Sensory Processes, The New Psychophysics", New York, Academic Press, 1974, page 153.

15. K.D. Kryter, "The Effects of Noise on Man", New York, Academic Press, 1970, page 265.

FIGURES

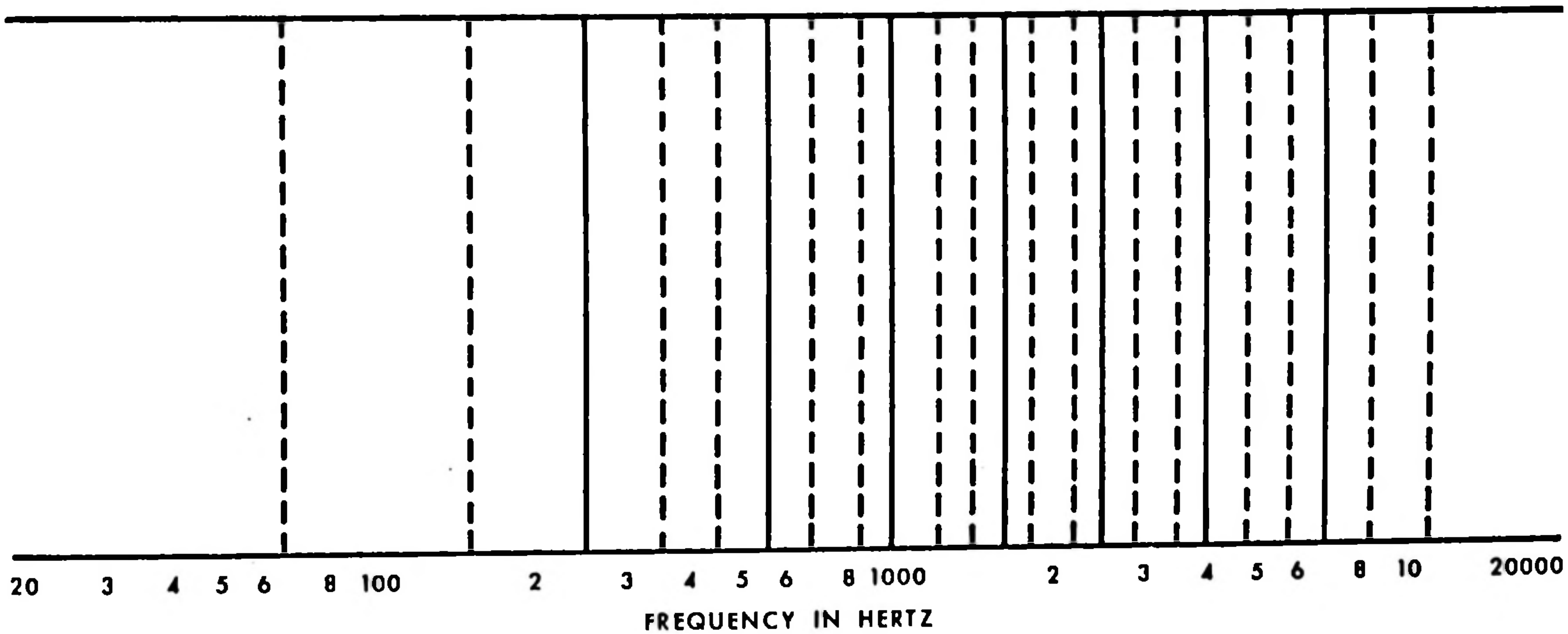


Figure 1. Critical Band Filters (Ideal)

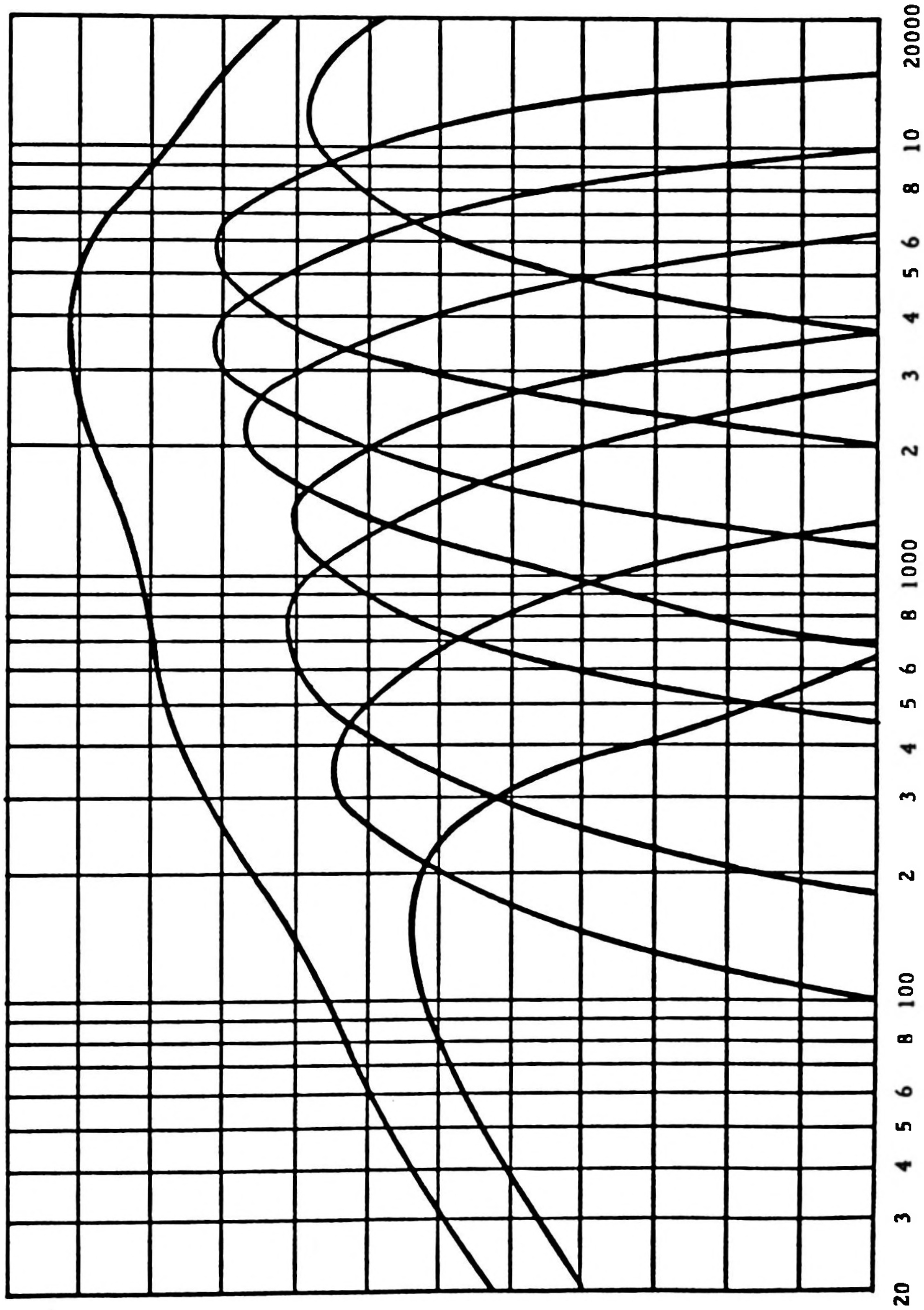


Figure 2: FREQUENCY IN HERTZ
Response of CBS Loudness Indicator

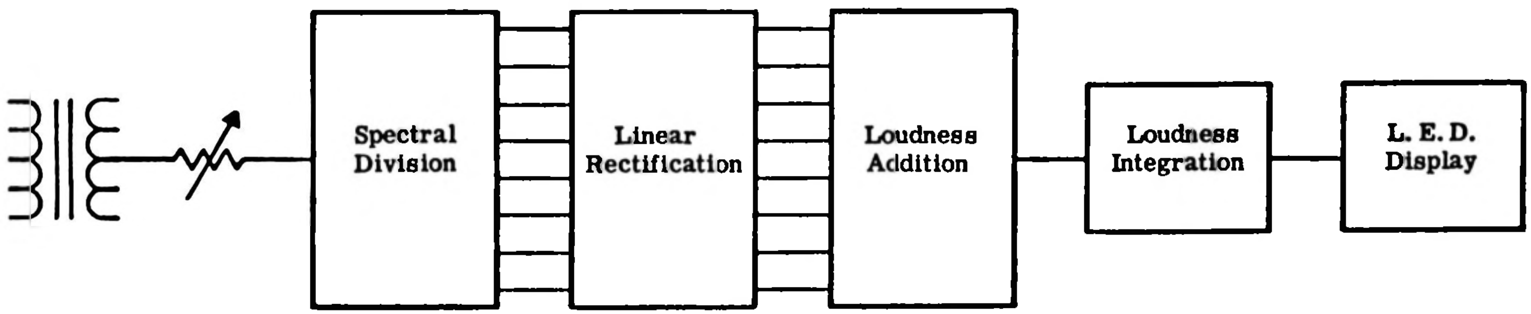


Figure 3. Block Diagram
of CBS Loudness Indicator

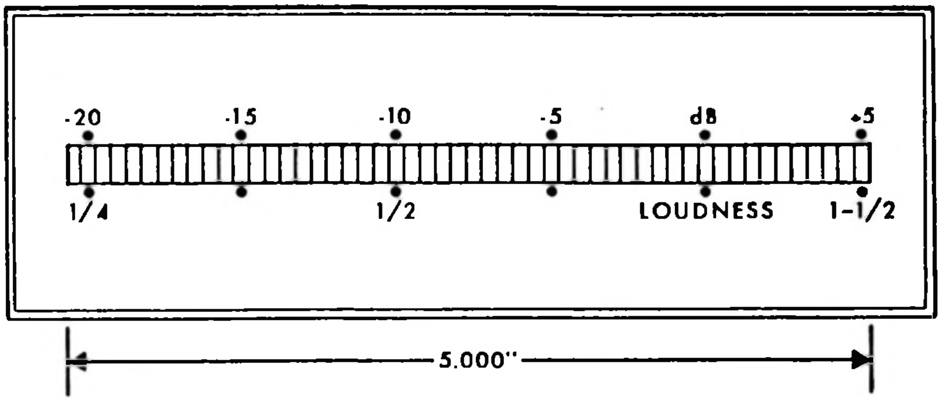


Figure 4. Display
on Front of CBS Loudness Indicator

CBS TECHNOLOGY CENTER

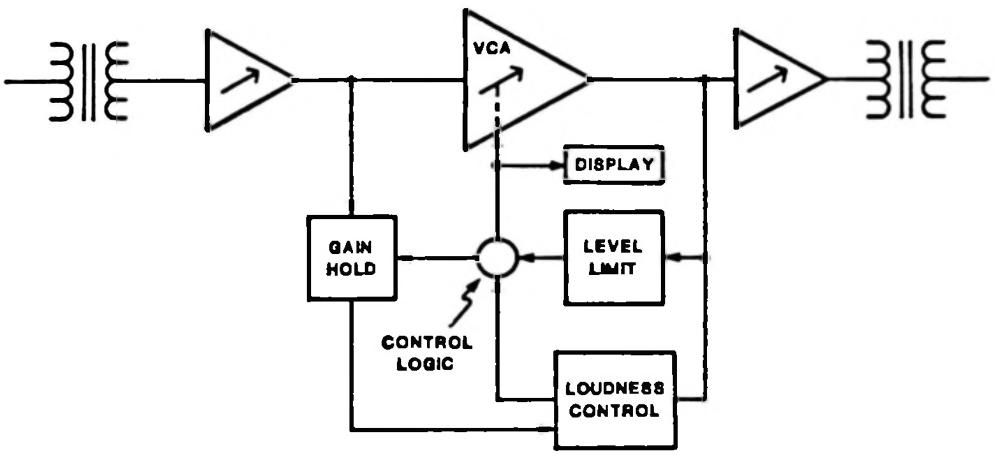


Fig. 5.
BLOCK DIAGRAM
CBS LOUDNESS CONTROLLER

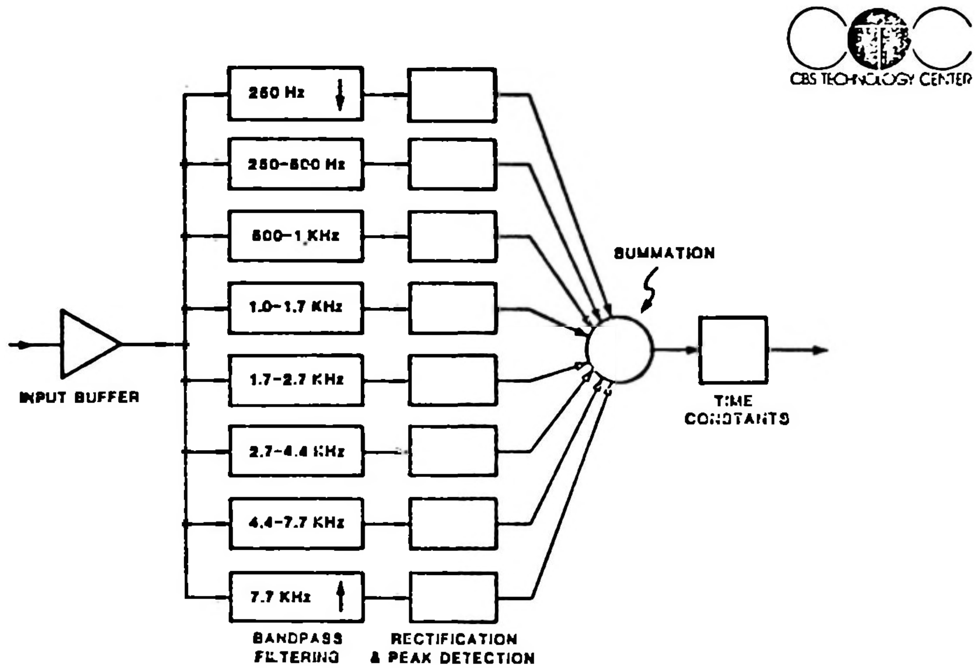
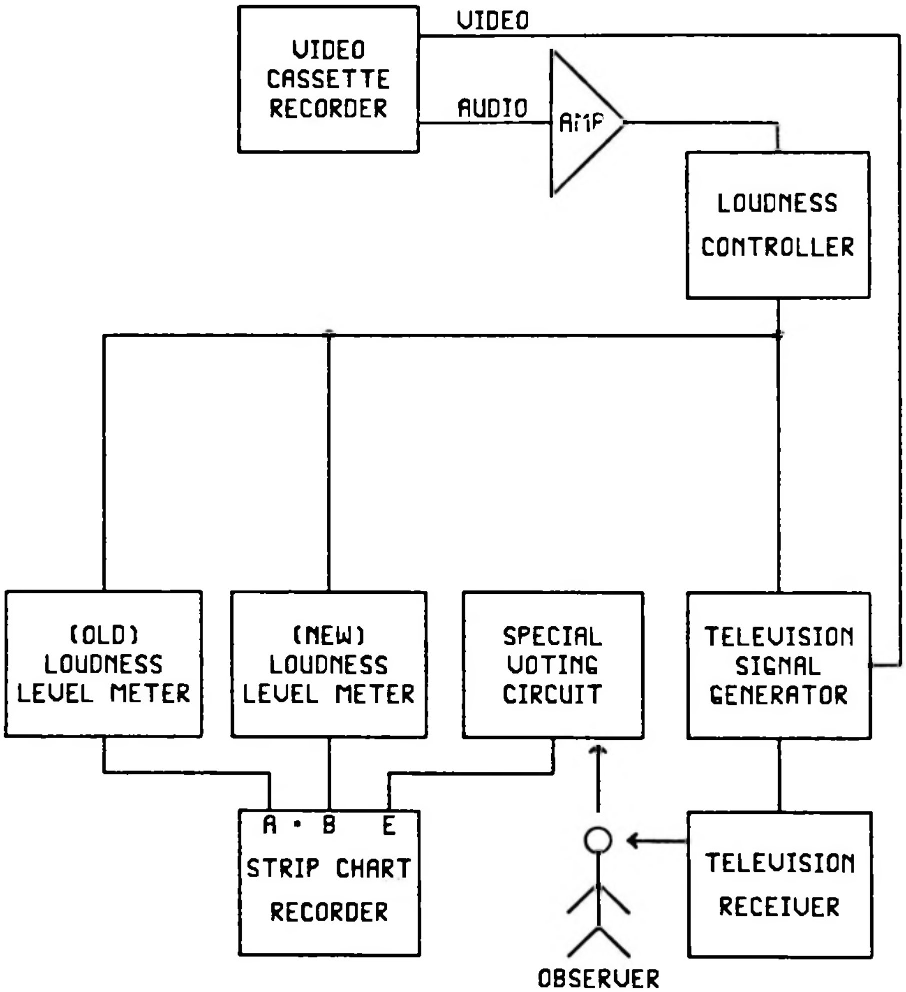


Fig. 6. Loudness Control Circuit Block Diagram
(See Figure 5)

APPENDICIES

APPENDIX 1

BLOCK DIAGRAM - TYPE I TESTS

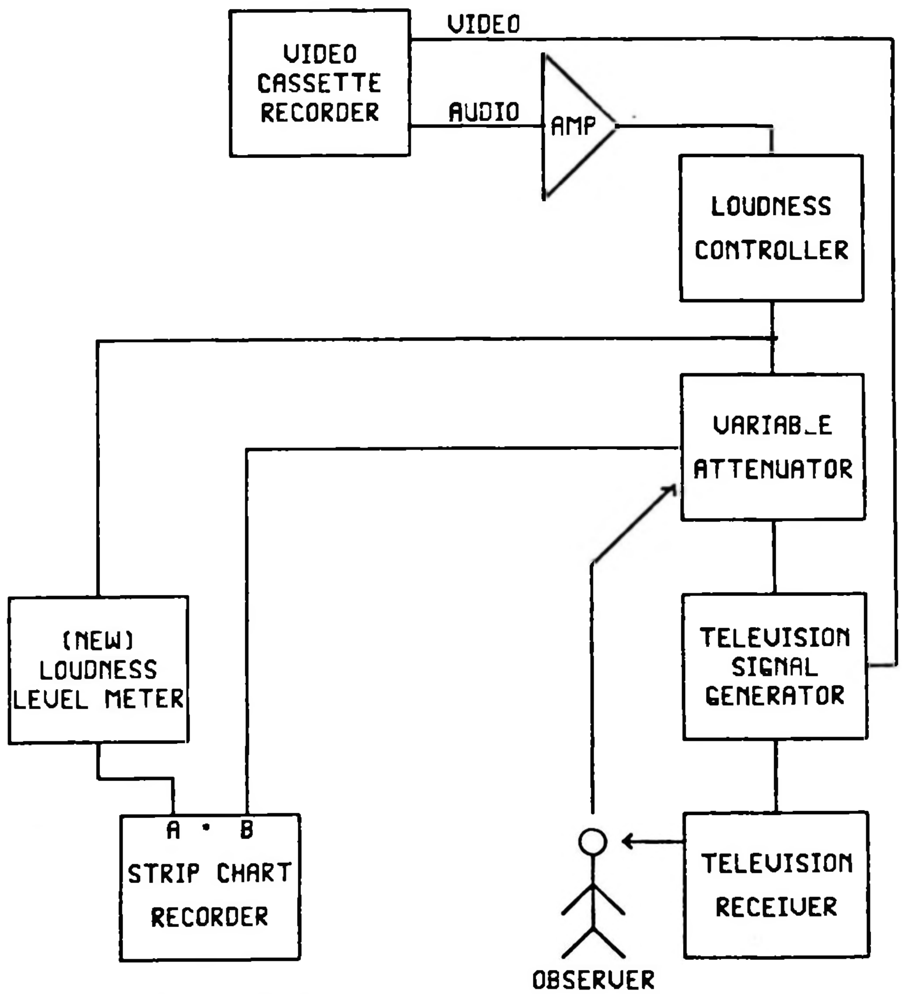


•STRIP CHART CHANNELS

NOTE - LEVEL CONTROL FUNCTION ENABLED OR DISABLED BY A FRONT PANEL SWITCH ON LOUDNESS CONTROLLER.

APPENDIX 2

BLOCK DIAGRAM - TYPE II TESTS



• STRIP CHART CHANNELS

NOTE - LEVEL CONTROL FUNCTION ENABLED OR DISABLED BY A FRONT PANEL SWITCH ON LOUDNESS CONTROLLER.

Appendix 3

INSTRUCTION SHEET

TYPE I TESTS
SUBJECTIVE LISTENING TEST

You are about to participate in a subjective listening test.

There is no right or wrong answer in this test. Set the volume control on the television to a level that is comfortable for your listening enjoyment. You will then hear program material and commercials. YOU are asked to make a judgement on the commericals' loudness based on the volume you set for the program material. In other words, if you feel the commercial is slightly louder than the program material momentarily depress, on the switch box in front of you: SLIGHTLY LOUDER.

If you feel the commercial is slightly softer than the program material momentarily depress: SLIGHTLY SOFTER

If you feel that the commercial is the same level as the program material then momentarily depress: PROGRAM LEVEL.

You have five choices available for each commercial you hear. These choices are, from left to right,

```
             SLIGHTLY     PROGRAM     SLIGHTLY
SOFT-------SOFTER-------LEVEL-------LOUDER-------LOUD
```

Please make your judgement during the commercial, and feel free to change that decision at any time during the commercial. The length of time that you depress the switch is not critical, one or two seconds is more than sufficient to record your selection.

THIS TEST WILL TAKE APPROXIMATELY 20 MINUTES.

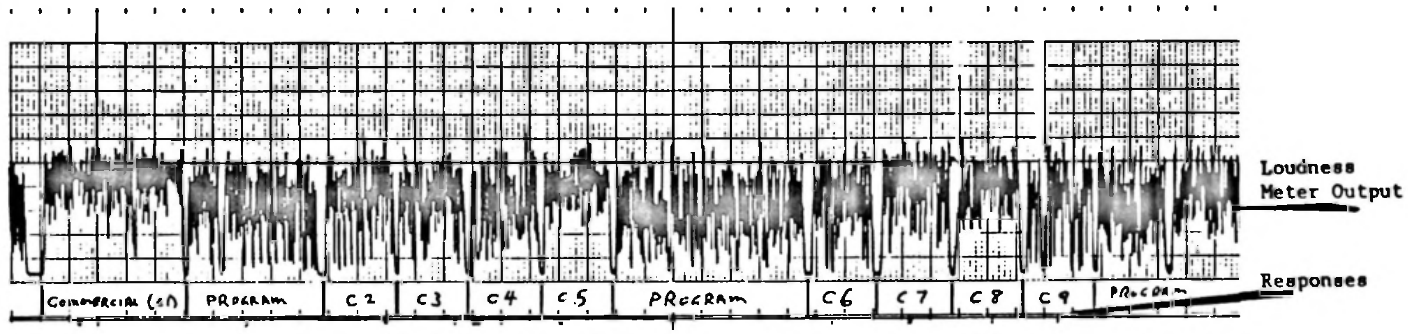
ARE YOU READY TO BEGIN?

Appendix 4

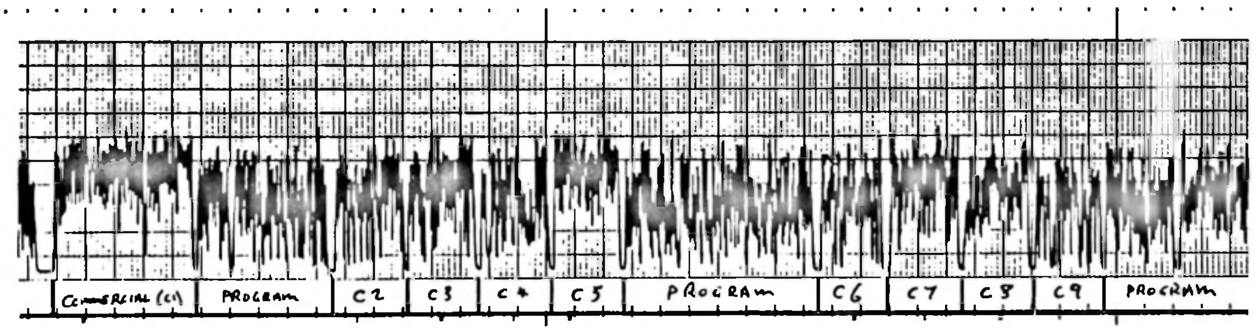
INSTRUCTION SHEET
Type II Tests

You are twice going to be shown a video tape of typical television programming and commercials. When the tape begins, adjust the volume control on the television set to a comfortable level for you. As the tape progresses, use only the remote volume control to keep the volume constant.

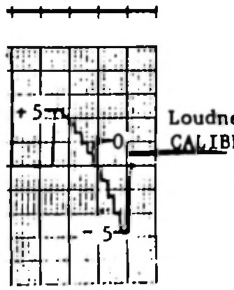
When viewing the tape the second time, do not attempt to remember your response from the first viewing, as the conditions of the audio may have changed. Always just respond with the remote control as you see fit to maintain constant volume.



CONTROLLER IN

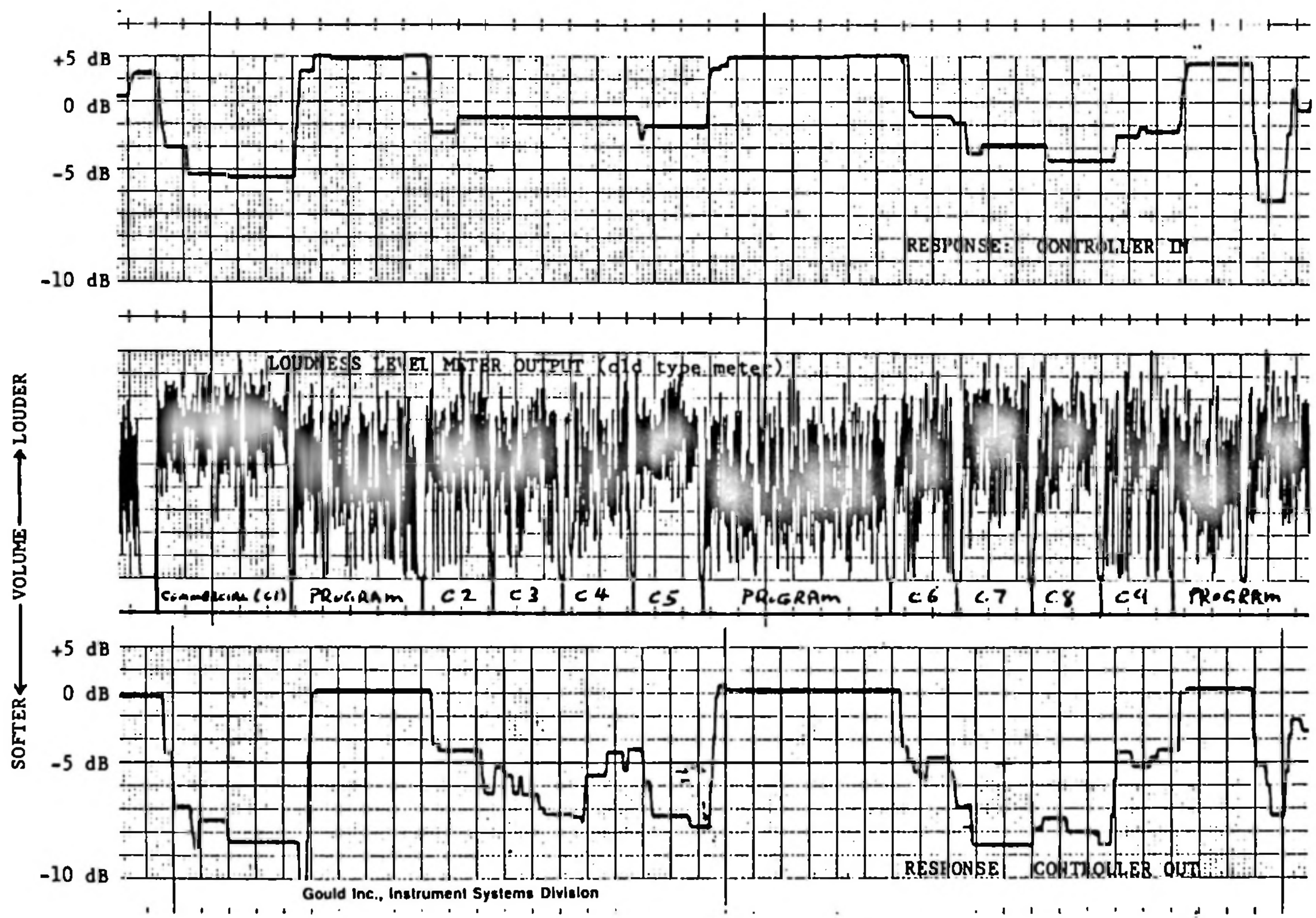


CONTROLLER OUT



Loudness Indicator
CALIBRATION [dB]

APPENDIX 5
Type I Tests-Strip Charts
Output of Loudness (new)
Indicator vs. Observer
Responses



APPENDIX 6
Type II Tests-Strip Charts
Observer Volume Control
Responses

APPENDIX 7: RAW DATA - TYPE I TESTS

COMMERCIAL	LOUDNESS INDICATOR IN	LOUDNESS INDICATOR OUT	PARTICIPANTS 1 IN	1 OUT	2 IN	2 OUT	3 IN	3 OUT	4 IN	4 OUT	5 IN	5 OUT
1	4-5	4-5	4	5	5	5	4.5	4.5	4	5	4	5
2	3-4	3-4	3	4	4.5	4.5	3	3.5	3	4	3*	4
3	3	3-4	4	4	4	4	3	3.5	3	5	4	3*
4	2-3	2-3	3	2	2	4	3	2.5	4	4	2	3
5	4-5	4-5	4	4	3	5	4	4	4	5	3	4
6	3	3	4	3	4	4	3	3.5	4	4	4	4
7	3-4	4-5	4	4	4	5	4	4	5	5	4	5
8	3	2-3	3	3	3	4	4	3.5	5	4	3	5
9	2-3	2-3	3	3	3	4	4	3	4	4	3	5
10	3-4	3-4	3	3	4	2	4	3	4	4	4	4
11	2-3	3	3	3	3	3	3	3	4	4	4	3.5
12	3-4	4-5	4	3	3.5	1.5	4	4	4	4	4	5
13	3-4	3-4	4	3	5	2	3	3	4	4	4	4
14	3-4	4-5	3	4	4	4	3	4	4	4	3.5	4
15	3-4	4-5	3	4	3	4	4	4	4	4	4	5
TOTALS	49.0	54.5	52.0	52.0	55.0	56.0	53.5	53.0	60.0	64.0	53.5	63.5

Note: Multiple level choices in a commercial were average to the nearest 0.5.
* No response = 3
Observation: Perfect level control would yield a total score of 45.0
(15 commercials X level 3)

Key:
1: Soft
2: Slightly Soft
3: Program Level
4: Slightly Loud
5: Loud

APPENDIX 7: RAW DATA - TYPE I TEST

COMMERCIAL	LOUDNESS INDICATOR IN	LOUDNESS INDICATOR OUT	6 IN	6 OUT	7 IN	7 OUT	8 IN	8 OUT	9 IN	9 OUT
1	4-5	4-5	4	4	4	4	4	4	4	4
2	3-4	3-4	4	4	3	3	3	3	3	3
3	3	3-4	4	4	4	4	3	4	3	4
4	2-3	2-3	3	4	4	3	3	3	3	3
5	4-5	4-5	3	4	4	4	4	4	3	4
6	3	3	3	4	3	4	3	4	4	3
7	3-4	4-5	4	4	4	4	4	3	3	4
8	3	2-3	3	4	4	4	3	4	3	4
9	2-3	2-3	3	4	4	4	3	3	3.5	4
10	3-4	3-4	3	3	4	3	3	3	3	3
11	2-3	3	3	3	3	3	2	2	3	3
12	3-4	4-5	3	4	4	3	4	3	3	3
13	3-4	3-4	3	3	4	4	3	3	3	3
14	3-4	4-5	4	4	4	3	3	4	3	4
15	3-4	4-5	4	4	4	4	4	4	3	4
TOTALS	49.0	54.5	51.0	57.0	57.0	54.0	49.0	51.0	47.5	53.0

Note: Multiple level choices in a commercial were average to the nearest 0.5.

Observation: Perfect level control would yield a total score of 45.0 (15 commercials X level 3)

Key:
1: Soft
2: Slightly Soft
3: Program Level
4: Slightly Loud
5: Loud

Appendix 8

RAW DATA - TYPE II TESTS

Subject	Controller In Range(dB)	Controller In dB	Controller Out Range(dB)	Controller Out dB
1	0.0 to - 7.5	7.5	+5.0 to - 7.5	12.5
2	-1.0 to - 9.0	8.0	-2.0 to -10.0	8.0
3	+5.0 to - 6.5	11.5	0.0 to -10.0	10.0
4	-7.0 to -10.0	3.0	-6.0 to - 9.0	3.0
5	-2.0 to - 7.5	5.5	-1.0 to - 8.0	7.0

Appendix 9

SUMMARY OF COMMERCIALS

1: SPONSOR: Gulf Oil Company
VISUAL : Overflight of pipeline and pumping station terminals
AURAL : Male announcer, normal delivery, high level background instrumental music

2: SPONSOR: Sominex
VISUAL : Business office
AURAL : Female speaker beginning, male announcer concluding, normal delivery, no music

3: SPONSOR: Triscuit
VISUAL : Triscuit on serving plate
AURAL : Vocal and instrumental music with male announcer, normal delivery

4: SPONSOR: Tums
VISUAL : Worker talking to "Mother Tums" on loading dock
AURAL : Conversation between worker and Mother Tums, no music

5: SPONSOR: FTD
VISUAL : Flowers being delivered
AURAL : Vocal and instrumental music, male announcer, normal delivery

6: SPONSOR: Jello Pudding
VISUAL : Families eating pudding
AURAL : Family conversations with male announcer, no music

7: SPONSOR: Bufferin
VISUAL : Person with headache
AURAL : Male speaker, serious delivery, tense instrumental music background

8: SPONSOR: Savings and Loans
VISUAL : "Happy" family scenes
AURAL : Group singing at beginning and end with male in middle, normal delivery over instrumental music

9: SPONSOR: Taster's Choice
VISUAL : Stacks of coffee cups
AURAL : Group singing at beginning and end with male announcer in middle, normal delivery over instrumental music

10: SPONSOR: King Pontiac City
VISUAL : New car showroon
AURAL : Male speaker, normal delivery, instrumental music background

11: SPONSOR: Roman Meal Bread
VISUAL : Man showing bread ingredients
AURAL : Male speaker, slow delivery, no music

12: SPONSOR: Pan American Airways
VISUAL : Travel scenes around the world
AURAL : Male speaker, slow delivery, concluding with sounds of happy people around the world, some singing, cheerleading, etc.

13: SPONSOR: King Pontiac City
VISUAL : New car showroom
AURAL : Male speaker, normal delivery, instrumental music background

14: SPONSOR: Wisk
VISUAL : Man, wife, and child
AURAL : Family conversation, young girl saying "ring aound the collar," male speaker, normal delivery, no music

15: SPONSOR: World of Wheels
VISUAL : Female speaker in tight shirt and hot pants, automobiles
AURAL : Female speaker, fast delivery, male announcer, normal delivery, no music

ACKNOWLEDGEMENTS

CBS Technology Center:	For development work in loudness methods and for providing hardware to be evaluated in the project.
Harris, Herbert C. :	Engineer, formerly with the FCC's Experimental Engineering Branch, for initial work in development of the project
Higgins, Lynne A. :	Secretary, FCC Experimental Engineering Branch, typing of report
Jones, Bronwyn L. :	Staff Scientist, CBS Technology Center, for support in development of the project and portions of the test describing CBS hardware
Torick, E.L. :	Director of Audio Systems Research, CBS Technology Center, for support in development of the project

Evaluation of the CBS Technology Center Loudness Indicator

Christopher P. Payne

ABSTRACT

The CBS Loudness Indicator provides indications of relative loudness in typical broadcast program material which very closely follows the human perception of loudness. With only one significant exception its accuracy appears to be generally within the test's experimental error of plus or minus 1 dB. An exception exists for string symphonic music. Listeners consistently found this music louder than did the loudness indicator. The testing did not provide an explanation of this anomaly.

If the sensing circuitry used in the loudness indicator is appropriately designed within an automatic transmitter modulation control device, the objectionable loudness variations presently experienced in broadcasting could be significantly reduced.

INTRODUCTION

The loudness of commercials has been a concern of the broadcasting industry and the Federal Communications Commission (FCC) for a number of years. Although many people believe that a radio or television station boosts the sound level when a commercial is broadcast, this is simply not the case; to the contrary, many stations and networks have specific operating practices which effectively reduce the transmission level of commercials in an attempt to reduce the number of loudness complaints.

The FCC investigated loudness in an attempt to provide a regulatory influence on the subject in several proceedings. In 1962, the Commission issued a Notice of Inquiry (Reference 1) seeking to identify the causes of complaints about annoying or objectionably loud commercials. In 1965, the Commission issued a Statement of Policy Concerning Loud Commercials in which broadcasters were cautioned to conform to certain specified operating practices thought to contribute to excessive loudness. In 1978, the FCC Field Operations Bureau conducted a study on

*This paper was written by Christopher Payne while he was employed at the National Association of Broadcasters. He is now employed by Motorola. The paper was presented at the 1983 NAB Convention by Michael Rau under the title "Broadcasting and Loudness: NAB Status Report".

loudness and reported its findings. Later, in July of 1979, another Notice of Inquiry was issued again seeking comments on how loudness could be better controlled: "...while the problem is not acute, it is certainly an irritant." The FCC disclosed they could not discover any evidence showing that stations were deliberately manipulating their signals to emphasize commercial messages.

The broadcasting industry has been attempting to find a satisfactory method to control the loudness of commercials. NAB has studied the problem, and, in the 1960s, produced a test tape and instruction booklet (s) to assist broadcasters in their operating procedures. In 1976, the NAB reactivated a subcommittee of its Engineering Advisory Committee (EAC) with a mandate to review the technology associated with loudness measurement and control.

International efforts have been organized under the International Radio Consultative Committee (CCIR). Several loudness tapes and indicating devices were considered within CCIR's Report 465-2(6). CCIR concluded "the determination of subjective loudness of programmes depends on too many imponderables..." and recommended that the particular working party studying the matter be dissolved. CCIR further recommended that a new study program begin which would investigate automatic methods to regulate loudness.

In 1982, these continuing efforts to measure and control loudness in broadcasting brought the subject to the attention of NAB's laboratory. In the mid-1960s, the CBS Technology Center (formerly CBS Laboratories) in Stamford, Connecticut, conducted experiments on loudness that included a review of the "equal loudness contours" developed by other researchers. The most relevant discovery during the experimentation, however, concerned loudness summation.

Conventional audio level meters, such as the ubiquitous VU meter and the sound level meter, did not respond according to loudness when two different signals were summed together. CBS discovered that if the signals were divided into certain frequency bands and subsequently added together, a much better correlation with perceived loudness could be attained. An instrument which would respond to relative loudness could now be constructed. Just this type of monitor was described in a paper published in 1967. The technology included empirically-confirmed time constants which caused the indicating meter to respond to loudness, with substantially greater accuracy than a standard VU meter. Comparisons of the new meter and the sound level meter are described in another paper published in 1971.

Although the new loudness meter was found to be a more accurate indicator of relative loudness than the sound level meter or the VU meter, the device itself, and a companion loudness controller, did not find wide acceptance in the broadcasting industry. Some controllers were sold -- and continue to be in use -- but the device is not widespread in the industry; and, broadcasters continue to receive loudness complaints.

In the meantime, work continued at CBS on refining previous techniques in loudness measurement and control. The previously constructed loudness monitor was modified and listener tests and experiments were again conducted. The results of the new work were reported in September of 1981.

Several of these loudness indicators were made available to NAB's EAC subcommittee on loudness to test at operating broadcast facilities. These meters were placed at various network and affiliate stations around the country for several months. A questionnaire was circulated with the meters and the results sent to NAB.

Many of these stations connected the meter either at the master control audio switcher output, or to a source of off-the-air programming and observed the loudness indicator as compared with a VU meter or modulation monitor. Some used the indicator for assistance in adjusting levels at a mixing console. Thus, the experience with the loudness indicator at the various broadcast operations was of a qualitative nature, and described its possible utility rather than affirming its accuracy in quantitative terms.

The expansion of the NAB Science and Technology Department to include the long planned for laboratory occurred in the summer of 1981. At that time, building space was made available, test equipment was delivered, and the laboratory began to take shape. The first project: investigate the loudness of program levels and verify the accuracy of the CBS Loudness Indicator.

THE CBS LOUDNESS INDICATOR

The CBS Loudness Indicator is designed to provide an indication of the relative loudness of electrical signals produced from acoustical sounds. In other words, the instrument is designed to connect to a conventional audio program circuit and provide an indication of relative loudness if the electrical signals are reproduced on an amplifier loudspeaker system. The development and validation of the instrument is described in a paper published in the SMPTE Journal. The CBS device differs from conventional audio level indicators in that it filters outputs according to an empirically derived formula. The sum is then integrated according to an experimentally derived time constant and the result indicated on a light emitting diode display. The design of the CBS indicator is modeled after the construction of the human ear.

THE NAB TESTS

Because the basic broadcasting objective is to improve the control of loudness variations, primarily in television programming, the NAB loudness investigation and loudness indicator validation applied primarily to television audio. Therefore, a television audio chain was simulated using an audio tape recorder, a CBS Model 4450A Audimax, a CBS Model 4110 Volumax (with de-emphasis), and a speaker system equalized according to a median acoutical response of 11 television receivers.

This arrangement of automatic level and peak controlling equipment is typical of the equipment in use by television stations today. A number of experiments were conducted which evaluated how effective this equipment was in controlling audio levels and modulation of an FM transmitter (with pre-emphasis). NAB concluded that the equipment performed its intended purpose. The Audimax did an excellent job of taking widely varying input levels and automatically adjusting the gain so that output levels had a uniform amplitude. The Volumax also did an excellent job of taking constant input levels and controlling the peak value of the audio output. especially maintaining the peak value of a pre-emphasized audio response. As might be expected, however, these units did not adequately control loudness variations in the programming. This is precisely the effect that caused broadcasters to receive complaints and desire a solution through improved equipment design.

The testing of the simulated TV audio chain produced considerable loudness variations when the audio contained widely different peak-to-average ratio (similar to the "crest factor" used in acoustics) and frequency spectrum. Various types of program material were sent through the program chain and the peak, average, and loudness characteristics were measured and recorded on a strip chart recorder. The loudest synthetically generated material which passed through the audio chain was noise, filtered to approximately a one third octave around 3 kHz and then severely clipped. The least loud program material was a male voice which had a very high peak-to-average ratio.

Experiments were conducted to determine if the least loud voice could be electronically altered to increase its loudness when sent through the program chain. The audio was highly compressed with a modern multiband limiter; and the frequency response was altered to increase the "presence" of frequencies centered around 3 kHz. The results show that the voice's loudness could be increased five dB over the unprocessed voice even if both signals had passed through the Audimax and Volumax.

A test tape of 13 segments of program material, each 30 seconds in duration, was recorded. One segment consisted of the clipped noise. All others were of voice and/or music. The tape was played through the program chain and heard through the equalized loudspeaker.

VALIDATION TEST

The technique for validating the accuracy of the CBS Loudness Indicator used a "loudness matching technique". This method asks a volunteer listener to compare the loudness of a 30 second segment of program, from the tape, with a constant level of USASI noise. The listener adjusts a volume control until he or she believes the two sounds are of equal loudness. The resulting listener-adjusted loudness program segment is then measured by the CBS Indicator and recorded on a chart recorder. If the listener and the indicator completely agree on the relative loudness of the different segments, the chart recording will show a constant loudness. Any difference between the listener's interpretation of loudness and the indicator's is the amount of disagreement.

THE EQUIPMENT

The Ampex 351 tape recorder feeds the Audimax which then drives the Volumax. The output of the Volumax passes through a one dB step calibrated attenuator and through the T-attenuator controlled by the volunteer listener. The output of the T-attenuator is fed to both the loudness indicator and to one position of a selector switch operated by the listener. The other input to the selector switch is connected to a General Radio 1382 noise generator. The output of the selector switch goes to the amplifier/equalizer and then to the loudspeaker system. The T-attenuator is mechanically coupled to a linear potentiometer which supplies a DC signal to the second channel of the chart recorder. The change in attenuation then can be read from the chart recording. The T-attenuator has a range of 20 dB with no shut-off. The DC output of the loudness indicator drives a DC log amplifier that in turn drives the first channel of the strip chart recorder. The result is a loudness indicator reading in dB on the chart recorder similar to the LED indicator on the loudness indicator's front panel.

THE TEST TAPE

If the T-attenuator is not changed during the tape playback, the loudness indicator will record its interpretation of the loudness variations which are passed through the Audimax/Volumax. The loudest segment, according to the loudness indicator, was found to be the "clipped" noise, and the softest material, the one labeled "Ben", a former NAB employee with a soft voice. The difference is about 10 dB. The volunteer listener, then, has to adjust the T-attenuator substantially if all the tape selections appear as loud as the USASI constant level noise. It should be emphasized that the listener is hearing the tape's audio after it passes through the Audimax/Volumax and through a speaker system adjusted to sound like a typical television receiver. In other words, the audio accurately represents an actual television transmission and reception system.

THE LISTENERS

Most of the listeners were volunteers from the NAB staff. Two were from the FCC, and two were from a local preparatory school. There was an equal male/female mix of people; ages ranged from sixteen to over sixty-five. A total of fourteen people participated in the tests.

THE PROCEDURE

Each listener sat in a chair and was provided with two controls: the T-attenuator and the selector switch. The loudspeaker was located about ten feet away. The reason for the test was explained to each listener. NAB was soliciting their opinion; the tests were not of the person, but of certain equipment. The function of the attentuator and selector switch was explained. The level of the USASI reference noise was set to a comfortable loudness (as preferred by the listener), generally about 70 dBa. Next, filtered noise was connected to the transmission system. The listener was asked to adjust the T-attenuator until the loudness of the filtered noise was identical to the USASI reference noise. The filtered noise was increased 5 dB, decreased 10 dB, and increased 5 dB to the original value; each time the listener attempted to balance the audio with the constant level USASI noise. This procedure was repeated until the listener could adjust the filtered noise to within about 1 dB of the same value for the various 5 dB steps. Several listeners were able to adjust the loudness variations well within a dB with very little effort. These listeners were considered competent to proceed to the next phase of the test: the loudness matching of program segments.

LOUDNESS MATCHING THE TAPE

The tape was played, and the listener was asked to adjust the loudness of each 30 second segment until it matched the USASI reference noise. The noise could be selected as often as necessary by the listener. Most listeners continually switched back and forth while adjusting the tape's loudness. The listener was informed that, if he was not satisfied at the end of 30 seconds that a loudness match had been achieved, the tape could be rewound and played again as many times as desired.

The ability to loudness "match" varies considerably among people. If a person could not readily find a loudness match, he was invited to make the tape sound deliberately too loud, then deliberately too soft, and, through successive approximations, find the equal-loudness point. This suggestion was of considerable help to some listeners who were then able to loudness match with an increased degree of certainty.

DATA ANALYSIS

Typical chart recordings show the segment between a particular listener and the loudness indicator. A line through the top of each chart recording segment represents the researcher's estimation of the loudness indicator's reading. The indicator is to be read in a manner similar to a VU meter; that is, not by taking the highest peak but by estimating the level of "peaks of frequent recurrence". The line is meant to represent this interpretation. The chart recording is analyzed first by marking the peaks of frequent recurrence, and then deriving the median value by applying a straight edge to the chart and moving it until half the data is above the edge and half is below. After the median line is drawn, the variation of each segment from the median is extracted and tabulated. If the observer's adjustment of equal loudness is the same as the readings of the loudness indicator, the data consists of a series of 13 lines through each program segment data aligned at the same value. Therefore, the reading of the loudness indicator above or below the median line represents the degree of disagreement between the listener and the loudness indicator for a particular segment, as compared to all segments.

Figure 1 is a tabulation of data taken from the chart recordings of the individual listener adjustments. Each listener is assigned a letter, A through N; the tape segments are listed at the left hand column. The median value taken from the chart recording is shown for each listener for every segment. To the far right, the median value for all listeners for each segment is shown.

INTERPRETATION OF THE RESULTS

The Loudness Indicator readings on the tape played through the Audimax and Volumax, unadjusted, and the final median adjusted values are graphed in Figure 2.

According to the loudness indicator, the material on the tape played through the Audimax and Volumax had loudness variations approaching 10 dB. For instance, "Ben Unprocessed", a first generation of an NAB employee whose voice had a high peak-to-average ratio was among the least loud, while a band of noise centered around 3 kHz and heavily clipped was indicated to be 10 dB louder. The group of listeners agreed with the loudness indicator; after adjustment to equal loudness, the indicator showed "Ben Unprocessed" to be minus one-half dB, and the band of noise at zero dB.

After observing many people go through many hours of loudness matching, obtaining some general impressions is inevitable. The technique appears to work quite well, and the accuracy is probably around plus or minus 1 dB for the group. Loudness matching is definitely an ability which is somewhere between a talent and an acquired skill. In general, nearly all people were included in the data if they were able to perform the "pre-test" of matching noise adequately. The results of the loudness matching should be interpreted with an uncertainty of about plus or minus 1 dB. This area of uncertainty includes nearly all the data except for one type of material, the symphonic music.

CONCLUSIONS

The CBS Loudness Indicator provides indications of relative loudness in typical broadcast program material which very closely follows the human perception of loudness. With only one significant exception, its accuracy appears to be generally within the test's experimental error of plus or minus 1 dB. An exception exists for string symphonic music. Listeners consistently found this music louder than did the loudness indicator. The testing did not provide an explanation for this anomaly.

If the sensing circuitry used in the loudness indicator is appropriately designed within an automatic transmitter modulation control device, objectionable loudness variations presently experienced in broadcasting could be significantly reduced.

\- Listeners A thru N -

Selection	A	B	C	D	E	F	G	H	I	J	K	L	M	N	Median
Ben	0	+2.5	+.5	-.5	-.5	-.5	0	-2	-1	+2	-1	-2	+1.5	-2.5	-.5
60 minutes	-2	+4.5	+2	+2.5	+1.0	+2	+.5	-1.5	+1.5	0	+3.5	+1	+3	0	+1.25
NAB	0	+3.5	+2.5	-0.5	-1.0	0	+2	0	-1	+.5	0	-.5	+1.5	-.5	0
Triumph	+.5	+2.5	+1.5	+2.5	-0.5	+1	+1	0	0	0	+1	+.5	+1.0	0	+0.75
Malox	-1	0	+2.5	+0.5	-1.5	-1.5	-1	0	-.5	0	0	0	-0.5	0	0
Rock	-1	-1	+2	+0.5	0	-1.5	-2.5	+2	-1	-.5	-2.5	+1.5	-2.0	-2.5	-1
Ben Proc.	+2.5	0	0	-2	0	+0.5	-1	+6	+1.5	-1.5	-2	+1.0	0	0	0
Symph	-1	-2	-4.5	0	-3.0	-3.5	-5.5	-1	-1	-6	-4.5	+1.5	-3.5	-3.0	-3
Bounce	+2.5	-1	-1.5	-1.5	0	0	-1	+3	0	-1.5	-1.5	0	-1.5	+1.5	-0.5
Soap	+1.5	-2	-2	-1.5	-1.0	-2	-3	-1.5	-1	-2.5	-2	-0.5	+.5	0	-1.5
Ben Rev.	+2	+.5	-1.5	-1	+4.5	+2	+1	0	+2	0	+1	+1.0	+1.0	0	+1
Storm	+3.5	-.5	-1	-2.5	+2.5	0	0	+3.5	0	0	+1	-1	-2.0	-3.0	0
Noise	-3	-.5	-1	-1	+5.5	+1	+1	+5	+1.5	+0.5	+6	-2	-4.0	-1.0	0

\- Tape Segments -

FIGURE I

Degree of Disagreement between Listeners and Loudness Indicator for a Variety of Program Segments.

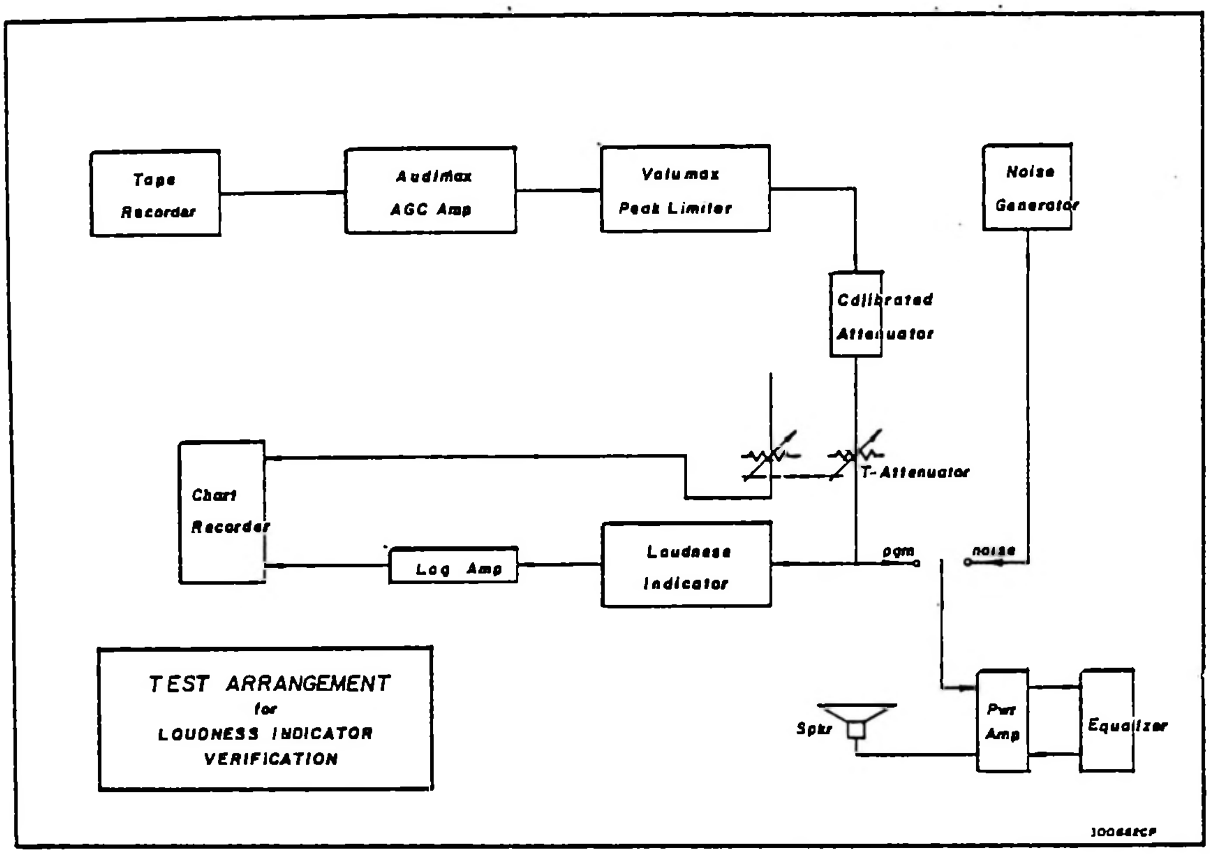
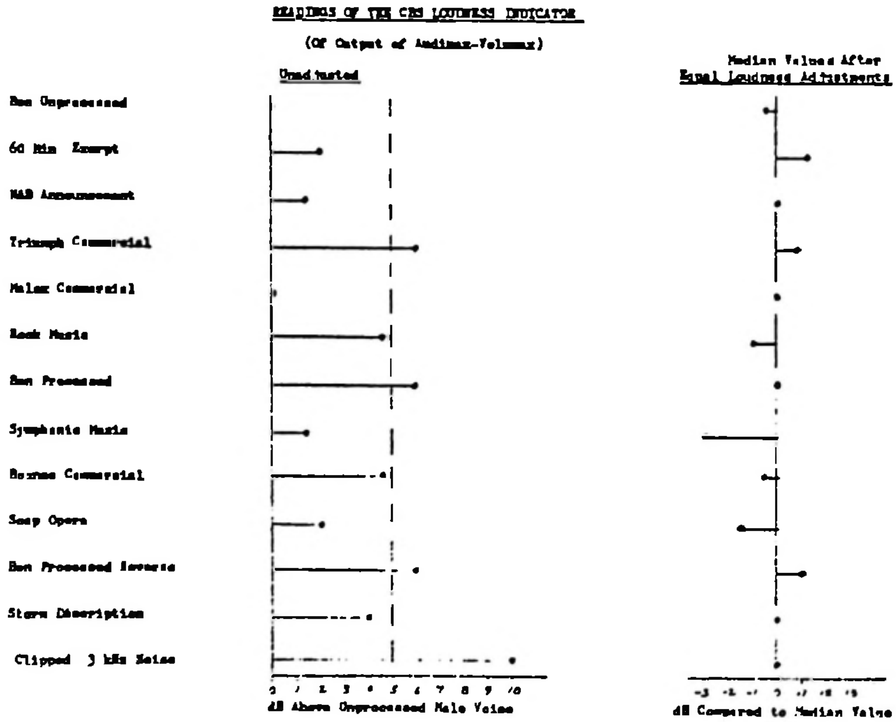

Tape Recorder
Audimax AGC Amp
Volumax Peak Limiter
Noise Generator
Calibrated Attenuator
T-Attenuator
Chart Recorder
Log Amp
Loudness Indicator
pgm
noise
Spkr
Pwr Amp
Equalizer
TEST ARRANGEMENT
for
LOUDNESS INDICATOR
VERIFICATION

FIGURE 2


READINGS OF THE CBS LOUDNESS INDICATOR
(Of Output of Audimax-Volumax)
Unadjusted
Median Values After
Equal Loudness Adjustments
Ben Unprocessed
60 Min Excerpt
NAB Announcement
Triumph Commercial
Malex Commercial
Rock Music
Ben Processed
Symphonic Music
Boxnes Commercial
Soap Opera
Ben Processed Reverse
Storm Description
Clipped 3 kHz Noise
0 1 2 3 4 5 6 7 8 9 10
dB Above Unprocessed Male Voice
dB Compared to Median Value

REFERENCES

1. "Notice of Inquiry", FCC Docket 14904, ("Amendment of Part 3 of the Commission Rules and Regulations to Eliminate Objectionable Loudness of Commercial Announcements and Commercial Continuity Over Standard, FM and Television Broadcast Stations"), Adopted December 17, 1982.

2. Report and Order FCC Docket 14904, Adopted July 9, 1965, Statement of Policy Concerning Loud Commercials, Public Notice FCC 65-618, July 12, 1965.

3. Notice of Inquiry, FCC Docket 79-168, ("Amendment of Part 73 of the Commission's Rules and Regulations to Eliminate Objectionable Loudness of Commercial Announcements and Commercial Continuity over AM, FM, and Television Broadcast Stations"). Adopted July 5, 1979, FCC 799-412.

4. W. Hassinger, "Evaluating Loud Commercials: An Experimental Approach" FCC/FOB 78/01, February 1978.

5. NAB Standard Loudness Reference Recording, Engineering Department National Association of Broadcasters, Washington, D.C. September 1967.

6. Recommendations and Reports of the CCIR, 1978, XIVth Plenary Assembly, Kyoto, 1978, Volume X, Report 465-2.

7. B.B. Bauer, and E.L. Torick, Researches in Loudness Measurement, IEEE Transactions on Audio and Electroacoustics, AU-14, September 1966.

8. B.B. Bauer, E.L. Torick, A.J. Rosenheck, and R.G. Allen, A Loudness-Level Monitor for Broadcasting, IEEE Transactions on Audio and Electroacoustics, Vol. AU-15, No. 4, December 1967.

9. B.B. Bauer, E.E. Torick, and R.G. Allen, The Measurement of Loudness Level, Journal of the Acoustical Society of America, Vol. 50, No. 2, August 1971.

10. B.L. Jones and E.L. Torick, A New Loudness Indicator For Use in Broadcasting, SMPTE Journal, September 1981, page 772.

11. M. Gourgon, Acoustic Performance of a Sample of Home Television Receivers, Development Report 2922-1, Operations Development Department, Canadian Broadcasting Corporation, January 1978.

List of Test Equipment

1. Ampex 351 Tape Recorder/Reproducer
2. CBS Laboratories Model 4450A Audimax
3. CBS Laboratories Model 4110 FM Volumax, modified for nonaural, with deemphasis
4. Daven attenuator set, 0 to 111 dB, 0, dB per step, 600 ohms
5. Shalco custom T-Attenuator, 600 ohms 1 dB per step, shaft extended out back
6. CBS Technology Center Loudness Indicator, prototype model
7. Hewlett Packard Dual Channel Chart Recorder, Model 7132A
8. (2) Technics SH-9020 Peak/Average Meter Unit
9. General Radio 1382 Random Noise Generator
10. Pioneer SX-780 Tuner/Amplifier
11. Altec 1653A Graphic Equalizer
12. Pioneer HPM-60 Loudspeaker System
13. Krohn-Hite 3100-R Variable Filter

ACKNOWLEDGEMENTS

The author wishes to acknowledge the cooperation of the CBS Technology Center, Stamford, Connecticut for the use of prototype equipment in these experiments. In particular, the author is indebted to Emil L. Torick, Bronwyn L. Jones, and John Byrne of the CBS Technology Center for their willing assistance and expert advice in psychoacoustic,testing and equipment design and operation. Additionally, Michael C. Rau of NAB's Science and Technology staff provided editorial assistance and guidance.

RF RADIATION IN BROADCASTING

JULES COHEN

JULES COHEN & ASSOCIATES, P. C.

WASHINGTON, D. C.

The United States system of broadcasting is often characterized as the best in the world. Of course, that boast is usually made by United States broadcasters, but the claim can be justified with little effort. One of the characteristics of the system deserving high praise is its diversity. Virtually all shades of opinion and the largest imaginable variety of entertainment formats are available to a substantial portion of our population. Even in sparsely settled areas, clear channel AM broadcast service and, at least, television translator service are likely to be available, even beyond the range of primary broadcast service.

That diversity of reception services necessarily requires a large number of transmitters. Some of those transmitters operate at substantial power levels. Including antenna gain, the effective radiated power of a directionalized AM station may be in the order of 200 kilowatts; low band VHF television and FM stations (Zone II) operate with effective radiated power of 100 kilowatts; high band VHF stations operate with effective radiated power of 316 kilowatts, and UHF television stations may operate with effective radiated powers of as high as 5000 kilowatts. Until approximately five years ago, the general public had little concern over the possible biological effects of the electromagnetic waves which made possible the broadcast services which entertained and informed. Then, in 1977, The Zapping of America: Microwaves, Their Deadly Risk, and the Cover-up, by Paul Brodeur, was published. In his book, Brodeur charged that a conspiracy by industry and the military was preventing the public from learning that the "electronic smog" being created by broadcasters, RF communications systems, military radar, and power lines affected their health adversely. The suggestion was made that experimental work hidden from the public showed that a large variety of human ailments could be attributed to nonionizing electromagnetic radiation. The effects included cancer, birth defects, blood disorders and various adverse behaviorial problems. The charges were documented by a very selective choice of published papers, in some cases, using information out of context.

The Zapping of America did not become a best seller, but the cry was picked up by writers such as Susan Schiefelbein, Jill Jonnes, and others. A Schiefelbein article which appeared in the September 15, 1979, issue of Saturday Review was entitled: The Invisible Threat and subtitled: The Stifled Story of Electric Waves. She continued the Brodeur theme that important data are being withheld from the public. The Reader's Digest contributed to the misinformation being disseminated by publishing a distorted account of the health effects on State Department employees of the bombardment with microwave radiation of the U. S. Embassy in Moscow by the Russians. A theme that is repeated in many of these articles is that three ambassadors to Russia subsequently died of cancer. What is not told is that, at the time these ambassadors served in Moscow, the ambassador's office in the Embassy was not subjected to microwave bombardment and the Russians never did direct any microwaves toward the ambassador's residence, which is a number of miles away. The fact is that the three ambassadors were elderly men with other health problems that had nothing whatsoever to do with microwave exposure.

As a result of these alarmist articles, not based on good scientific evidence, many citizens became concerned that exposure to radio waves may cause physical harm. Troubles soon developed for holders of FCC construction permits in the fields of broadcasting and communications. In most cases, the construction of a broadcast or microwave relay facility requires some form of variance from a local ordinance. Usually, either a special-use permit is necessary or a variance in the zoning ordinance height limitation is necessary. In any case, not only is a local building permit required, but also a hearing before a board of zoning appeals and perhaps a town board as well for the necessary authority to construct.

The board hearings have been used as the vehicle for resisting the construction of facilities on the grounds of creation of a hazard to the public health. When experts in the field of bioeffects of nonionizing electromagnetic radiation testify, they are often asked what proof is available that the normally very low levels of energy to which the public is exposed will not cause harm. The answer is usually that experimental work at moderate levels show no evidence of harm and that no mechanism has been conceived which would provide a rationale for adverse health effects at very low levels. But the negative cannot be proven in absolute terms. All that can be said accurately is that no hazard has yet been observed and the probability of finding hazard is extremely low. Such an answer is generally not acceptable to the lay public.

In view of the alarmist literature with respect to RF fields, the public would like to have some reassurance, supplied by an unbiased source that their exposure to particular electromagnetic signals will not affect them adversely. The engineering and biological experts brought in to tesitify in hearings are, of course, paid by the proponent. Unfortunately, many members of the public will not accept the objectivity of engineers and scientists under such circumstances. These people want the reassurance to come from a source such as the Environmental Protection Agency. However, the wheels of that Agency grind very slowly, indeed. Despite the fact that work within the Agency directed toward a protection guide has been going on for several years, the first publicized step, an Advance Notice of Proposed Recommendations, did not appear until the December 23, 1982, publication in the Federal Register. Absent the availability of federal protection guides and local expertise for setting standards, some localities (the towns of Onondaga and Pompey, in New York, and Multnomah County in Oregon) adopted moratoria on the construction of any new radio facilities.

If the broadcasters of this country are not to be thwarted in their attempts to broaden and improve reception, the public must somehow be made to understand that only in the most unusual of circumstances, the ambient field strength from broadcast facilities does not constitute a health hazard. Furthermore, protection guides based on sound scientific principles are needed from the EPA. Additionally, an ongoing research program is necessary to fill the gaps in our knowledge and determine the mechanisms for interaction of electromagnetic fields and living tissue. Unfortunately, the present administration is eliminating funds for research and is generally opposed to anything smacking of "regulation." No funding has been made available for continuance of the Electromagnetic Radiation Management Advisory Council (ERMAC), the EPA is requesting no fiscal year 1984 funds for the radio-frequency and microwave radiation program at the agency's Health Effects Research Laboratory in Research Traingle Park, North Carolina, the group at the National Institute for Occupational Safety and Health which had been working on a criteria document has been dispersed, and the National Telecommunications and Information Administration (NTIA) has given up its role in coordinating federal research in RF/MW bioeffects research.

If the Executive Branch cannot be convinced of the need for protection guides and a continuing research program, perhaps the Legislative Branch can be induced to require the continuance, and even acceleration, of research and standards development.

IMPROVING PICTURE QUALITY IN

THE NTSC SYSTEM

Kerns H. Powers

RCA Laboratories

Princeton, New Jersey 08540

Recent interest in high definition television stems from two very different and distinct applications. One is the possibility of a new television service to the consumer to satisfy the needs of a high quality, wide screen display in the home. The other is its application to motion picture production through the process of electronic cinematography and post-production. First of all let me say I am very bullish on the prognosis for the development of a higher definition system for the electronic cinematography application.

But now let me speak to the first application from the vantage point of a receiver manufacturer. When I speak of HDTV, I am tacitly implying a new system with a wide aspect ratio and resolution approximately double today's TV in both the horizontal and vertical directions. It is clear that major investments would be required over several years duration to introduce into the home a high definition television service. During that same period, there will be investments in higher resolution cameras, video tape recorders, and digital editing suites for the electronic cinematography application. It is not quite so clear that there will also be the large investments in transmitters, cable equipments, and direct broadcast satellites or other transmission systems that would be required to get high definition from its point of origin to the receivers. On the receive end, however, most of the investments that would be required for HDTV in the receivers themselves and especially in the picture tubes and the projection displays will be made to improve the quality even for NTSC signal displays. In other words, most of the new technology required for HDTV will be developed independently of any new high definition consumer service.

Since the introduction of color transmission with the NTSC system in the early 1950s, there has been a steady and gradual but continuous improvement in the quality of the displays in the home. Those early monochrome receivers had a bandwidth of 2.7 MHz. There was little incentive to make the receivers any better than that at the time because the common carrier transmission facilities were less than 3 MHz bandwidth. But that situation did change with color and gradually over the years the quality of all television equipments have continued to improve in performance.

The introduction of digital technology into the television studios, and the handling of component rather than composite signals, will substantially improve the quality of the NTSC signal transmitted in the last link from the VHF or UHF transmitter to the receiver. The receiver must keep pace with this improved quality and it will.

The introduction of the comb filter in television receivers about three years ago is the latest example of the gradual improvement in picture quality from evolving technology. Now the concept of the comb filter has been around for many years but it is only recently that integrated circuit technology has reached the stage that a low-cost comb-filter color decoder could be manufactured in large quantities and included in the television set. In the past, the color decoder in TV sets has contained a low pass filter to separate the luminance signal from the color subcarrier components thus limiting the horizontal resolution to the equivalent of about 3 MHz. Also because of the imperfect separation of luminance and chrominance, the cross-talk between these components has resulted in the artifacts known as cross-color and cross-luminance which are manifest in the color fringes that appear on the herringbone jackets of the anchormen and the dot crawl evident on the edges of the characters during rolling credits. The comb filter has substantially eliminated both of these artifacts while at the same time, opening up the luminance bandwidth to the full 4.2 MHz. Although the comb filter causes a reduction of vertical color resolution and still suffers a bit of cross-color at 45^{o} detail regions, this is a small penalty to pay for the substantial improvement in the sharpness and the overall quality of the displays in modern day receivers.

Now, if one can see ahead to the time when integrated circuit technology makes possible the promised "frame store on a chip," the TV receiver can include that fabulous color decoder called the frame comb which would totally eliminate, at least in stationary portions of the picture, all of these cross-color effects without loss of vertical color resolution. There will still be some cross-color and loss of resolution on moving objects in the picture but moving objects require less definition anyway as the eye cannot perceive definition in motion.

But with a frame store in the receiver, not only can we eliminate cross-color and dot crawl, but we can wipe out vertically every flickering artifact that gives today's television its classical "look." Let's look at today's 525-line system as we would approach close-up to a wide screen display and let's examine not the whole picture but only a center section to expand the things we would see, as shown in the monochrome test pattern of Figure 1(a).

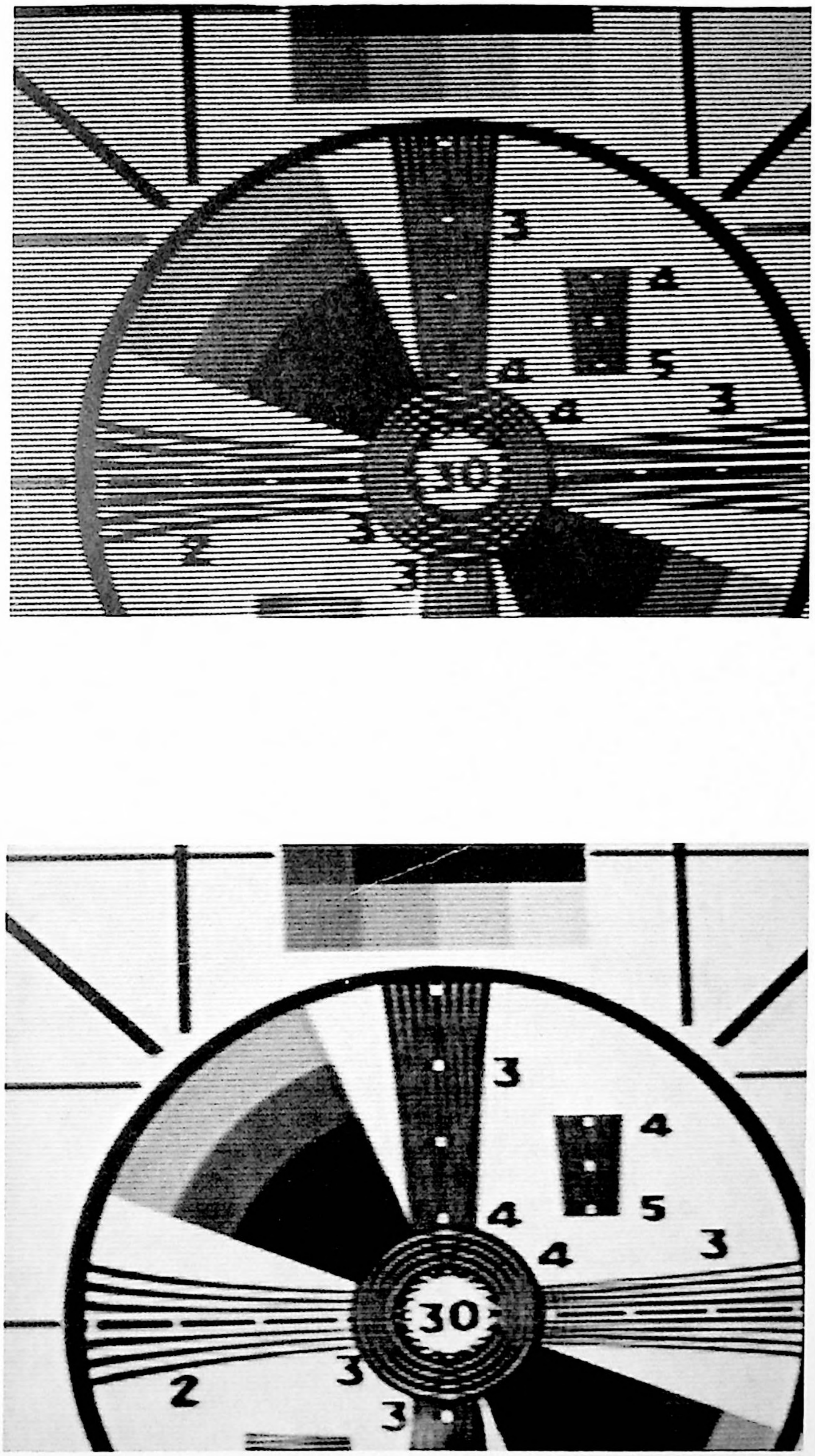
One of the obvious things you see when you display a single field is the raster of black lines. That is a result of the interlace technique. On the successive fields of a given frame, those black lines will be overwritten by the second field and supposedly they disappear. But at a distance from the display of one, two, or even three times the picture height, you will be able to see very clearly these black lines, and furthermore the interlacing does not eliminate them, it only subjectively sets them in motion and in fact any time you see an object in motion vertically you see the line-pairing effect of black lines moving vertically up or down. This effect is especially apparent when the signal-to-noise ratio is high and when you are in close to the screen.

Figure 1(b) is the same test pattern exposed for 1/30 of a second or for a full frame. Now it is obvious that if we have a frame store in the receiver that has a full frame of information stored, it can be read out of the frame store at a higher line rate. It can be read out at double the line rate or triple the line rate. More importantly, it can be read out non-interlaced in a progressive display so that one full frame is displayed in a 60th of a second rather than a 30th, and we will have eliminated completely this crawling black raster. We have not increased the resolution to do this, but we will have substantially improved the subjective quality of the picture.

Now, let's take a look at another phenomenon that appears on the test pattern of Figure 1(a). Let me call your attention to the horizontal wedge of high vertical detail. A single field contains only 262½ lines but on the horizontal wedge you see all kinds of funny things going on. This is what we call vertical aliasing. Vertical aliasing results from having a higher resolution picture than 262½ vertical samples can adequately resolve. But on examination of Figure 1(b), the aliasing apparently vanishes when a full frame of 525-lines is scanned. In our interlaced system, the aliasing components on successive fields do not really cancel but rather, they flicker at a 30 Hz rate. TV engineers know very well the "busyness" in the horizontal wedge of the test pattern that I call the flickering vertical alias. If we store a full frame in a frame store and then read out that frame in a 60th of a second, we would not see any of this busyness in that horizontal wedge. So a lot of the moiree that we see on herringbone suits and striped ties on the television screen will also be eliminated if we use a frame store to eliminate the flickering alias effects associated with interlace scanning.

I have so far been talking about still pictures but in pictures that have motion content, another artifact occurs that will not be eliminated by the frame store. This is the serrations that appear on the edges of moving objects. The serrated edges result from the fact that two successive fields are not of the same scene because the object has moved in that 60th of a second from one field to the next. So any object in motion will have line breakup on vertical edges because the edges are displayed with only half the vertical resolution. To eliminate the serrated edges of a moving object will require a smart signal processor in the receiver which adaptively interpolates between lines rather than selecting alternate lines from the frame store. Alternatively, this effect could be eliminated at the camera with a shutter to guarantee that two fields are scanned from the identical image. Professor Broder Wendland of the University of Dortmund in West Germany has given a number of talks before the SMPTE on methods that can be used in a frame store associated with a high definition camera to eliminate vertical aliasing, serrated edge effects, and other artifacts within the constraints of a 525-line 30 frame per second transmission medium. Professor Wendland has proposed to scan the HDTV camera at double rate (say, 1050 lines), sample field-quincunx horizontally, perform two dimensional digital filtering of the image in a frame store to achieve a 2-D Kell factor approaching unity (rather than the value of 0.6 or so typical of the camera optics MTF) thence select alternate samples and alternate lines for encoding into a standard 525-line interlaced NTSC signal. This procedure, coupled with the post-filtering made possible by the frame store in the receiver, is alleged to achieve an increase in effective resolution by a factor of 40-50 percent, without requiring <u>any</u> increase in transmission bandwidth.

These improvements in the NTSC system will take place, not from consumer demand, but from competitive pressures in the marketplace, whether a wideband HDTV service is introduced or not. The basic open question is whether the increased resolution of HDTV over these artifact-free NTSC displays will have sufficient perceived value to justify the investment in television plant required to accommodate the increased bandwidth.



(a) One Field

(b) Two Fields

Figure 1. Vertical Aliasing

NTSC High Quality Television Receiver

Richard S. Prodan

Philips Laboratories

Briarcliff Manor, New York

Abstract

The introduction of video picture memories at modest cost will provide a means of improving picture quality with greatly reduced artifacts in consumer television receivers. The subjective improvement of the television picture can be accomplished with a combination of digital signal processing and memory in the receiver without the need for changing the transmission standard or preprocessing at the transmitter. This full compatability with existing standards (NTSC, PAL, SECAM) allows the introduction of such improved receivers without problems. An NTSC high quality receiver system that has been constructed and is under development and evaluation at our laboratories is described.

Receiver Design Objectives

The major objective in high quality television is improvement of the subjective appearance of the displayed image. A substantial improvement within the existing standards seems possible, since present receivers do not fully utilize all of the available information in the transmitted signal. It is necessary to remove the artifacts present in conventional receivers to see an improved picture.

Degradation of image quality due to noise, area flicker, line flicker, interlace effects, cross color and cross luminance should be substantially reduced. Some details of the system that has been implemented for the NTSC standard to accomplish these objectives are presented.

Noise and Cross Color Reduction

The receiver processing structure is based on a first order recursive digital low pass filter. The basic configuration of the filter is shown in Figure 1. The incoming video signal is reduced by a factor K and added to the previously stored information in the memory reduced by a factor 1-K to provide the displayed output signal and new input information to the memory. This action averages the incoming video signal with previous frames on an individual picture element basis. It can be shown that the frequency response of the filter is a comb filter with spacing between the teeth equal to the picture frequency of 30 Hz and spectral width of the teeth determined by the value of K. The amplitude at the peaks remains at a constant value of unity. This characteristic is therefore matched to the spectrum of the video signal while random noise is largely rejected. This type of filter has been used for noise reduction in professional equipment without apparent loss of picture information. This results from the temporal rather than spatial filter characteristic requiring one complete picture memory as the filter delay element.

In the case of stationary pictures, no adverse effects appear. However when there is motion in the picture, objectionable lag occurs in the moving portions of the picture that increases with increased filtering. This undesirable effect can be overcome by using a motion detector that changes the filter coefficient on individual picture elements in the moving portions of the picture. This necessitates sampling the video signal at the same locations in each frame.

If the luminance signal is filtered, the interleaved components of the color subcarrier fall in between the peaks of the comb response thereby greatly reducing cross color. Similarly, when filtering the two baseband chrominance signals, the cross luminance components are located at the nulls of the filter response and are reduced. The filtering performed is essentially an averaging process over many pictures with the most recent pictures contributing the most to the average. The averaging of television frames leaves the zero average value of the cross components, as well as the zero average value of the random noise at the output.

The receiver signal processor block diagram is shown in Figure 2. The low frequency and high frequency portions of the luminance signal are applied to separate filters with coefficients K1 and K2. The two chrominance signals are applied to two filters with the same coefficient K3. The values of the three coefficients are determined by a motion detector. The recombined luminance and two color difference signals are converted to different scanning standards before display. Using a display scanning standard differing from the transmitted one allows the reduction of display generated artifacts such as area flicker, line flicker, and interlace effects. The techniques used to remove these undesirable effects due to the display scanning process is the subject of the following discussion.

Display Scan Conversion

The signals are converted to double frequency for display by various methods to be described to eliminate area and line flicker effects. The conversions fall into two categories. The first method displays a 525 line interlaced picture at double line frequency and double field frequency. The second method displays a 525 line sequential picture at double line frequency and normal field frequency. Both methods require twice the transmitted signal bandwidth at the display. The first method eliminates area flicker due to the 120 Hz field frequency and can also reduce line flicker. The second method eliminates line flicker and reduces loss of vertical resolution due to interlace effects.

The 120 Hz field rate interlaced method is divided into two strategies. Let the first field of the picture be called field A and the second field B. There are two different sequences to display these fields at double field frequency, namely AABB or ABAB. For the displayed information to be spatially correct, the AABB sequence must be interlaced in the odd field position for the first two displayed fields AA and switched to the even field position for the next two fields BB. The information is displayed in proper time sequence (field B always follows field A) so that no motion effects are observed. The interlacing of the fields is at the original 30 Hz rate so that line flicker is still present and only large area flicker is eliminated. The ABAB field sequence is spatially correct for proper interlace, so line flicker will occur at twice the original rate. This greatly reduces the visibility of the line flicker, as well as the area flicker. The information is displaced from the normal time sequence (field A is displayed after field B) which creates a disturbing discontinuous display of motion in the picture. Motion detection is necessary to switch between these two field sequences on moving parts of the picture. The scanning beam must also be adaptively displaced for correctly displaying the information spatially. This leads to a hard decision between stationary and moving picture elements.

The 60 Hz frame rate sequential method is similarly divided into two strategies. The sequential display is generated by interpolating the missing scan lines in a field and displaying the resulting 525 line sequence at twice the standard line rate. Two strategies for the interpolation are employed. The first uses the previous field in memory to supply all missing lines in the present field. This eliminates line flicker completely by displaying the same sequential picture during each field period. Furthermore there is no loss of spatial resolution since no averaging of adjacent picture elements is required. This previous field interpolation technique works well for stationary pictures, but produces disturbing serrations on moving images due to the separation in time between the two fields at the source. If motion is present, the other strategy is to interpolate a missing line in the present field by using the two adjacent lines from the same field. The time difference between two lines in the same field is much smaller resulting in negligible serration of moving edges. The averaging of information in adjacent lines results in a loss of vertical resolution. This necessitates motion adaption between the two interpolation strategies. In the sequential display conversion, a smooth transition between both field and line interpolation is possible by combining a proportional amount of each as a function of the amount of motion present in a picture element, since either source of interpolated information is always spatially correct.

Conclusion

An NTSC high quality television receiver with subjectively improved picture quality and reduced artifacts has been described. The system is compatible with the existing standard. The signal processor and display have been realized and are undergoing further development and evaluation. Preliminary results indicate that motion detection is very important for optimum performance of the system and will be the subject of further experimental work.

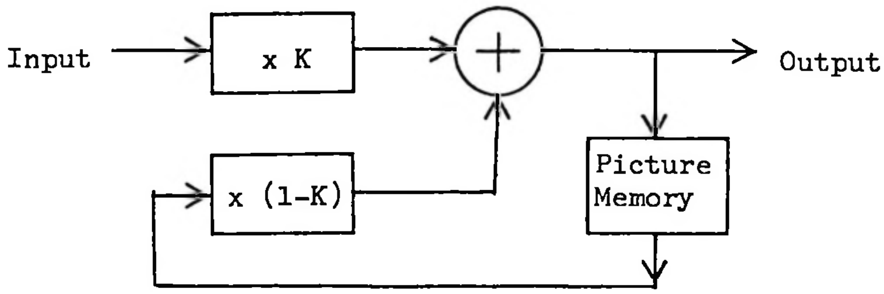


Figure 1 : Recursive Filter

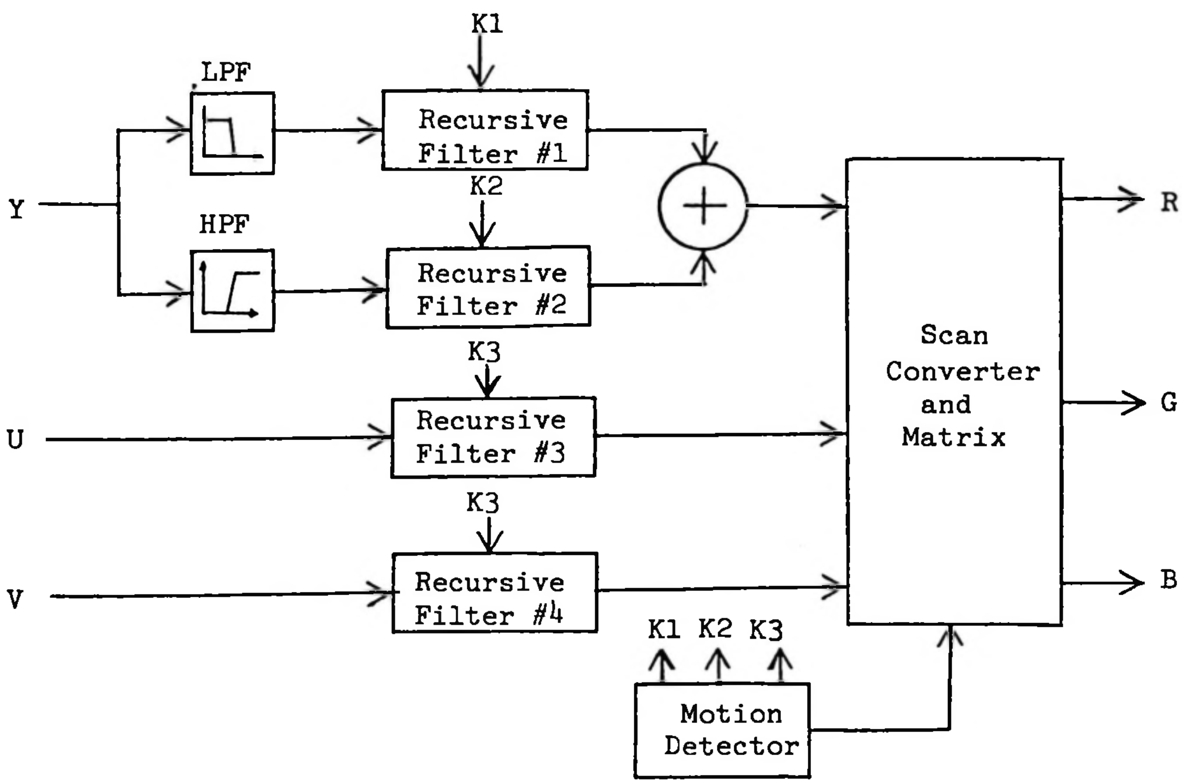


Figure 2 : Receiver Signal Processor

COMPATIBLE TRANSMISSION OF HIGH DEFINITION TELEVISION USING BANDWIDTH REDUCTION

W. E. Glenn, Ph.D., Karen Glenn, Ph.D., R. L. Dhein, I. C. Abrahams

New York Institute of Technology Science and Technology Research Center

Dania, Florida

The National Television Systems Committee compatible color transmission system was developed by taking advantage of psychophysical phenomena that were known to be present in human color vision. It was known that the eye could not perceive chromaticity information at as high a resolution as luminance information. Consequently the high resolution chromaticity information did not need to be transmitted in color television images. This resulted in a reduction in the bandwidth required for color television transmission of approximately two to one. In order to make it economically feasible to introduce color without losing the audience that watched black and white sets, the system was wisely chosen to be compatible with black and white reception.

Recent psychophysical measurements[1] of the spatial and temporal response of the eye, particularly in the presence of masking effects due to image motion, have made further extensions in the understanding of human vision. This information, combined with the availability of frame stores, has made it possible to use the same philosophy in extending color transmission to high definition as was used for NTSC color. In the proposed transmission system, information that the eye cannot perceive has been removed from the transmitted signal. In a monochrome simulation designed to take advantage of the results of the psychophysical measurements we have been able to reduce the transmitted bandwidth by approximately three to one without detectable artifacts in normal program material and test targets. If it is desired to produce a high definition television image with twice the vertical resolution and four times the horizontal resolution (wide screen), a bandwidth of eight channels would normally be required. Using the bandwidth reduction system we have devised, this image can be transmitted using one standard channel that an unmodified receiver can reproduce as a 525-line image, plus an additional channel that contains the information required to reconstruct the full HDTV image on a receiver using the special decoding. Even though this system requires a frame store in the receiver, it has other bonuses. The display scan rate can be in any format desired. It can be either interlaced or progressive. The retrace blanking can be whatever is desired and the vertical scan frequency can be chosen to be above normal rates.

Principle of Operation

It has been found that in vision the transmission of information from the retina to the brain is carried by two types of neurons. One type, referred to as "transient" neurons, carries rapidly changing information at low resolution from the retina to the brain. The receptors for these neurons are more numerous in peripheral vision than in central vision (in the fovea). This receptor-neuron combination has a time constant of about 50 milliseconds. They are most effective in detecting motion and flicker but have much lower spatial resolution than the limiting resolution of the eye. Receptor-neuron combinations located primarily in central vision (the fovea) have a much longer time constant and are referred to as "sustained" neurons. This receptor-neuron combination has a time constant of approximately 200 milliseconds and detects high resolution information in central vision. Because of a long time constant they are not excited efficiently in rapidly changing parts of the image.

There is also a masking phenomenom that is important in moving images. Transient excitation produces suppression of reception of detail information by sustained neurons for a period of about 200 milliseconds. Because of this suppression and the long time constant of the sustained neurons in the presence of a rapid brightness change caused by motion, it is not necessary to update detail information at the same frame rate as low resolution information. Consequently in the new system the normal 525-line transmission which is updated at 30 frames per second is used to provide the low resolution rapidly changing portion of the image. The detail information which is transmitted in an additional spectrum is updated at about 5 frames per second. The combination of both of these signals can be contained in the spectrum allocation of less than two channels for an HDTV image which has eight times the information content if transmitted at 30 frames per second.

Psychophysical Measurements

Extensive psychophysical measurements have been made to expand on the information available in the literature to provide sufficient data to design the transmission system. We measured the threshold of contrast sensitivity for monochrome luminance gratings and isoluminance chromaticity gratings at a number of temporal frequencies over the spatial frequency range of importance in television at normal viewing distances. This provided a series of modulation transfer functions of the eye at these spatial and temporal frequencies. This not only provides some of the information required in determining the time constants of the eye to both luminance and chrominance - it also provides an accurate re-evaluation of the desired proportions in resolution required for luminance and the primary color difference signals. As a result we would recommend somewhat better chromaticity resolution than is present in the NTSC transmissions. The high resolution portion of the color difference signals can be transmitted at lower frame rates in the same way as the detail luminance signal.

An abridged version of the steady state contrast sensitivity curves of the eye from this study is shown in Fig. 1. This figure shows response of the eye for luminance and for isoluminance complementary color pairs on lines on the CIE diagram which go through luminance C white and the three NTSC primary colors. The color curves have been normalized so that the peak sensitivity of all four curves is equal. This normalization makes the minimum perceptible color difference equal in contrast to the minimum perceptible luminance contrast for all three primaries at low spatial frequencies. The curves are terminated at the spatial frequency at which color is no longer perceived. Gratings above this spatial frequency are still visible but appear to be monochrome at all color contrasts. From these curves and termination points, it is evident that the color difference signals should be about half the luminance bandwidth for R-Y and one-quarter that bandwidth for B-Y.

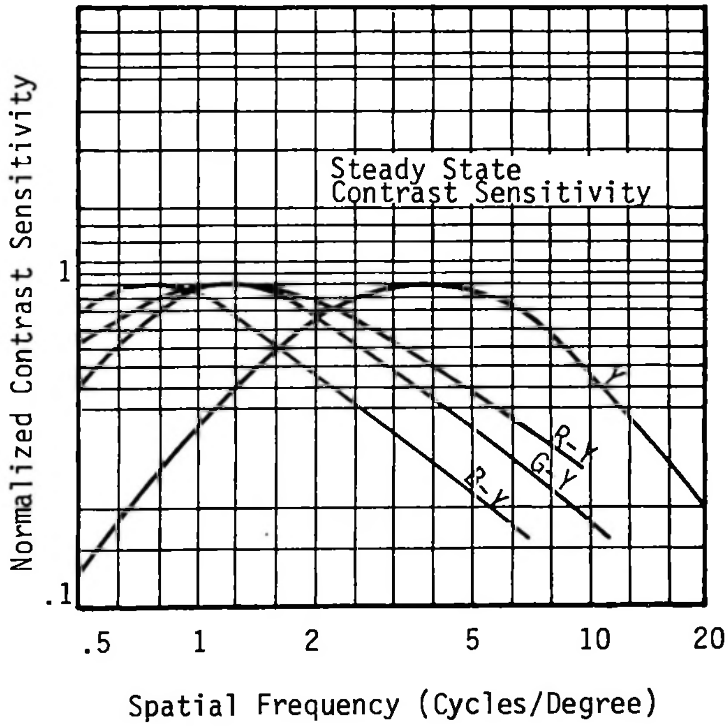


Fig. 1. Steady State Contrast Sensitivity

A second series of measurements were made to determine the perception of spatial frequencies as a function of duration. We used durations that were 1, 2, 4, 8, 16, 32, 64 and 128 fields (17 milliseconds per field.)

Figure 2. shows the relative sensitivity of the eye to luminosity gratings as a function of the duration of presentation of the grating. This is shown at .6, 4 and 12 cycles per degree. At low spatial frequencies in both luminance and chrominance, the eye adapts in time to the grating making it appear less visible. For short duration presentations at about .1 second, these spatial frequencies are greatly enhanced in visibility compared with the steady state visibility of the grating. For very short presentations, all spatial frequencies are suppressed.

Figure 3. shows the corresponding curves for isoluminance chromaticity gratings as a function of duration at .2, .9 and 2.7 cycles per degree.

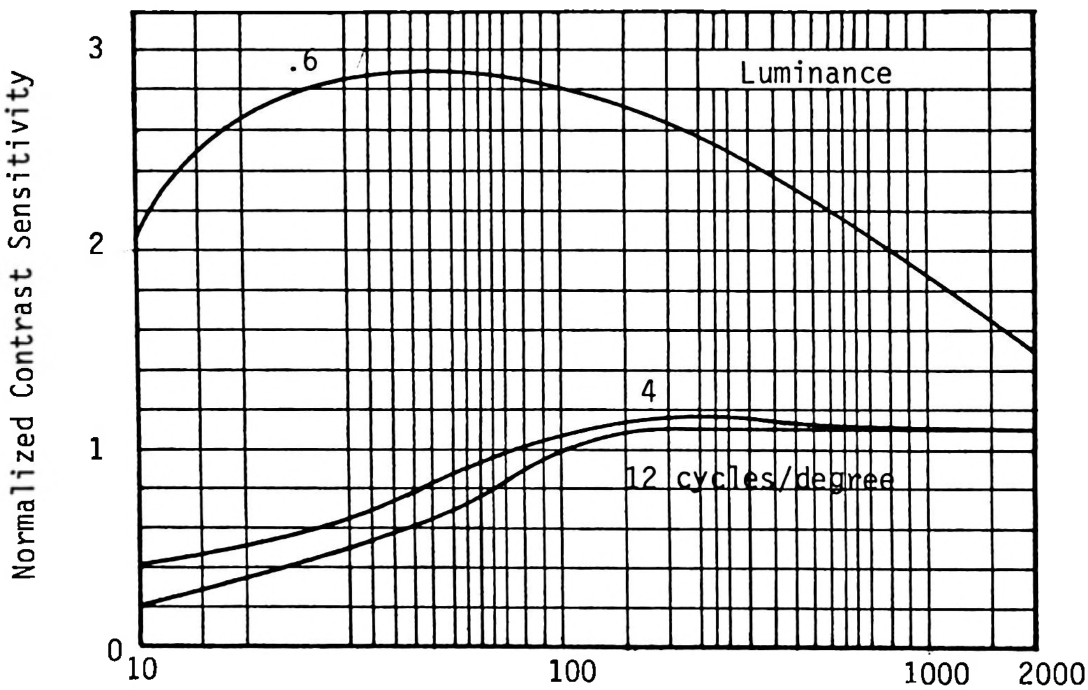


Fig. 2. Luminance sensitivity as a function of duration.

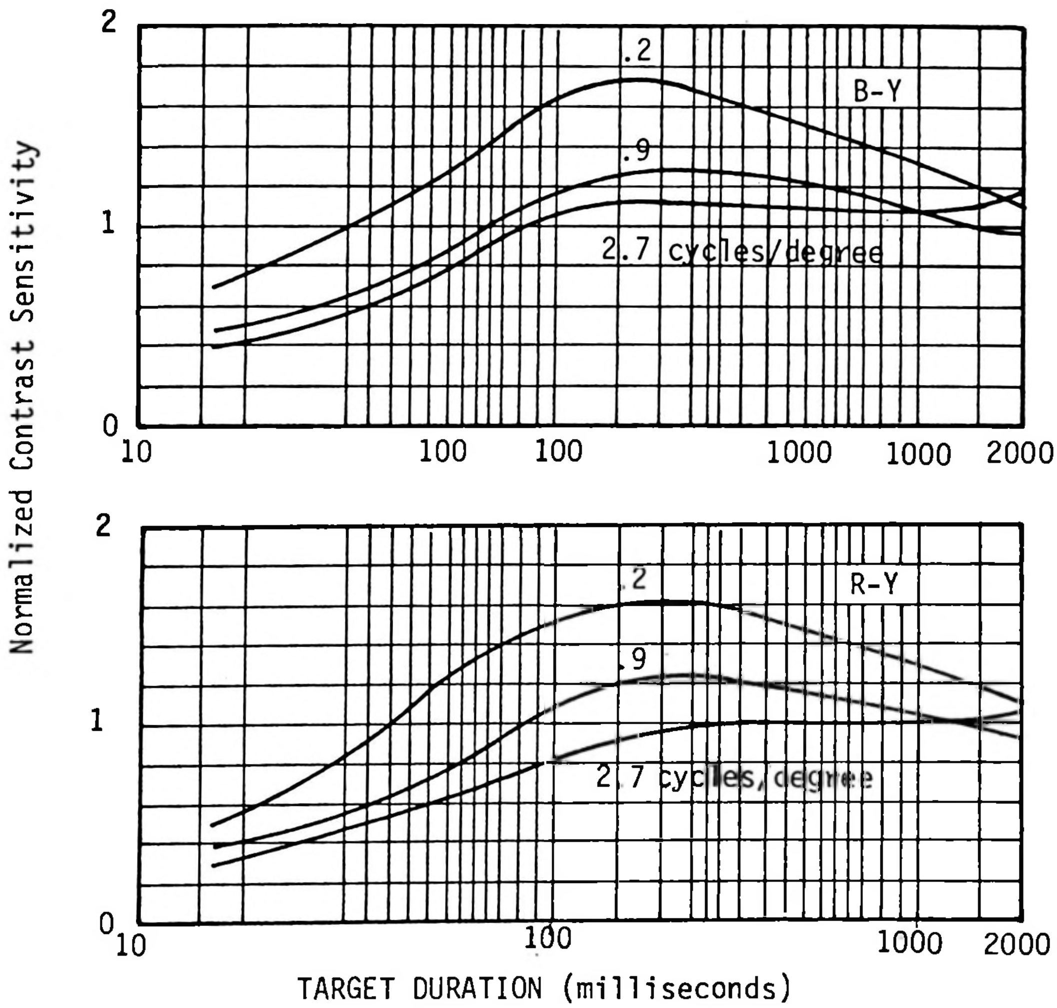


Fig. 3. Chrominance sensitivity as a function of duration.

In the presence of motion there is generally a low spatial frequency brightness change which suppresses reception for a period of time. The suppression of the grating (target) actually precedes the mask for a short time. We presented the target before the mask (backward masking) as well as after the mask (forward masking). We measured the degree of suppression by a .1 second mask using off-the-air television as signal during the mask. The spatial frequencies (target grating) were presented for various intervals either immediately preceding or immediately succeeding the mask.

Figure 4. shows a plot of the relative response of the eye to these stimuli as the grating duration was changed either immediately preceding or succeeding the mask. This experiment combines the effects of duration (Figs. 2. and 3.) and masking. It represents what would happen to the sensitivity of the eye in detail information in the scene that was just uncovered by a moving object in the image.

Figure 5. shows the corresponding curve to Fig. 4. for isoluminance chromaticity gratings. As one can see, the suppression is significant in excess of 200 millisecond period above 4 cycles per degree in luminance and above .9 cycle per degree in chrominance. The masking is primarily forward masking for both luminance and chrominance. The suppression of low spatial frequency luminance information in the target occurs only for about 50 milliseconds before or after the mask. From this one would conclude that a frame rate of about 20 frames per second is necessary to depict motion at low spatial frequencies. Motion picture film people have known this for a long time.

Experimental Results

To implement a simulation of this system, a processor was constructed using a frame store in both the transmitter and receiver. The simulation was scaled down by a factor of two below the HDTV version. It had 525 lines in the high resolution display and half that for the "compatible" signal. This results in the same spatial and temporal frequencies in the simulation but half the field of view of the HDTV system. A camera was progressively scanned and scan converted by the frame store to interlaced scan for display. The "compatible" transmission consisted of transmitting the sum of clusters of pixels that were two pixels high and two pixels wide. In the final, wide-screen version this would probably be two high by four wide. This transmission is made at the normal interlaced thirty frames per second. Detail luminance difference signals are transmitted at five frames per second. From these signals the detail information is reconstructed within the cluster. The resulting image updates the average of the pixels within the cluster to the correct value every 1/30 of a second. It uniquely reconstructs the value of each pixel within the cluster to its proper brightness (while maintaining the rapidly updated sum) with an update interval that never exceeds 1/5 second but averages 1/10 second.

We are now in the process of extending the system to color using psychophysical data collected on the spatial, temporal and masking effects in vision of isoluminance chromaticity gratings.

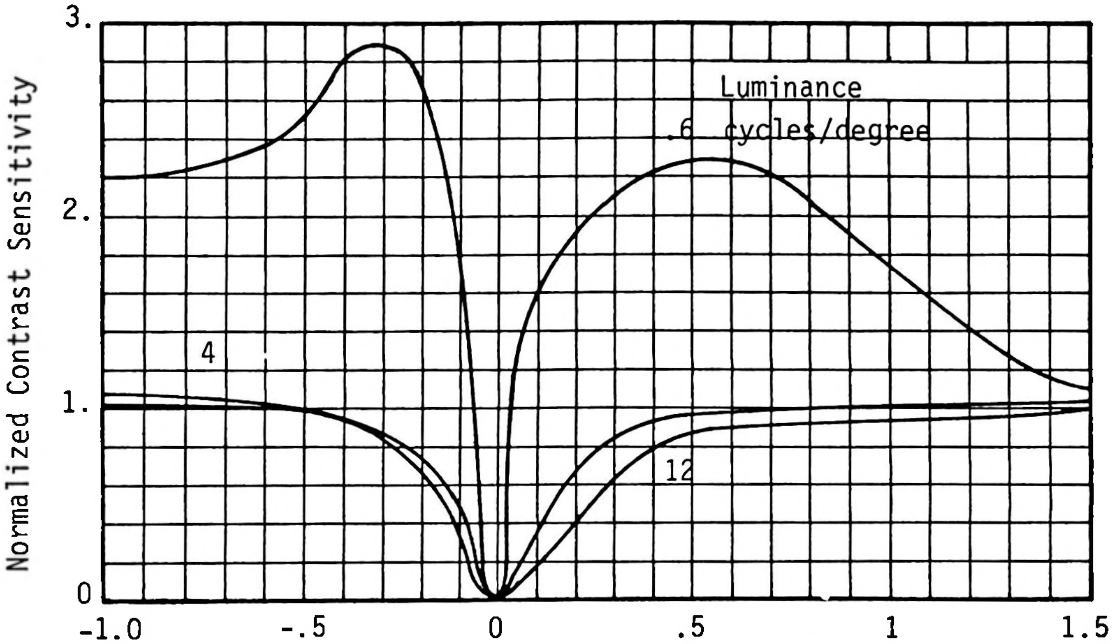


Fig. 4. Luminance masking by a .1 sec mask.

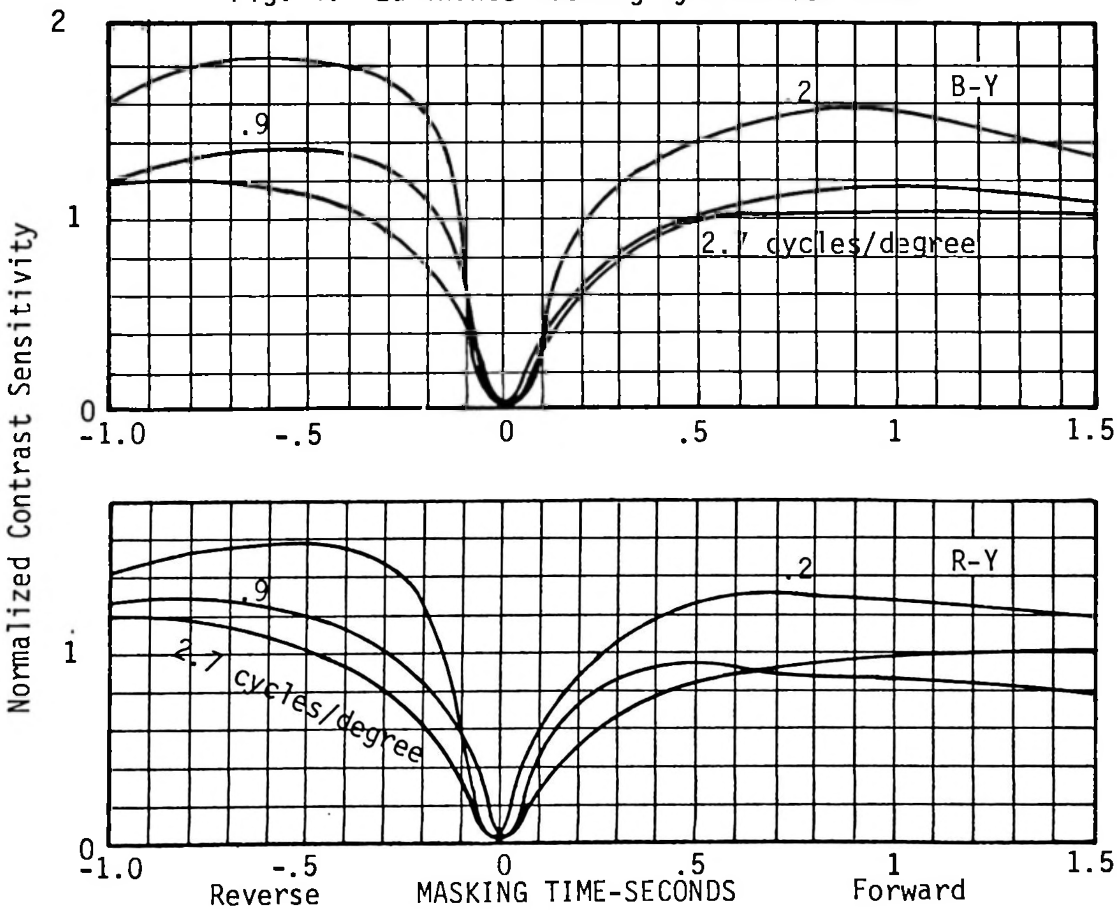


Fig. 5. Chrominance masking by a .1 sec mask.

Proposed System

The proposed transmission system will use normal NTSC standards for the compatible transmission and a time-base corrected component system for the detail luminance and chrominance transmission. Cross-color effects can be eliminated by restricting the bandwidth of the compatible signal that is used in the HDTV receiver to eliminate the information between 2.5 and 3 megahertz where cross-color is most severe. The component detail transmission will reconstruct luminance above 2.5 megahertz and chrominance above .5 megahertz.

The information transmitted at 2.5 megahertz for luminance appears at about 4 cycles per degree on an HDTV screen when viewed at a distance of three screen heights. The corresponding NTSC color carrier can resolve color up to about one cycle per degree under similar conditions. It is clear that by using appropriate time delays in the signal processing at the transmitter, the luminance and color difference signals above these frequencies can be masked in the presence of low resolution brightness changes caused by motion.

As explained previously, the color resolution need not be more than half the luminance resolution either horizontally or vertically. Since the base band transmission in NTSC standards has 525-line vertical resolution in color, this is adequate vertical color resolution for HDTV transmission. The color transmission needs to provide additional color detail information in the horizontal direction only. It is recommended that the compatible (30 frames per second) NTSC color be used as the low resolution color information and that the horizontal detail color information be transmitted with two slow scan analog components with one-half the luminance detail at a color axis of about 100^o on a vectorscope and one-quarter this detail at a color axis of 10^o. These axes were derived from our measurements and agree very closely with those recommended by NHK.[2] A detailed description of the psychophysical measurements will be published later.

We have made psychophysical measurements pertaining to compatibility of aspect ratio between the standard receiver and an HDTV receiver. A preliminary report on this study has been submitted to the Society of Motion Picture and Television Engineers HDTV Subcommittee on Psychophysics.

As in the early phases of development of NTSC color, there still needs to be a considerable amount of work done to assure that artifacts are minimized for all types of subject matter. However, we feel that there is a firm technical foundation for establishing a compatible transmission system with a considerable reduction in bandwidth. The system based on the data we have collected would have a base band transmission which is a completely standard NTSC transmission. And additional 4 to 5 megahertz is required to transmit the detail of both luminance and chrominance using analog component transmission. Information can be transmitted during most of the blanking interval in this channel. The resulting image would have 1000-line by 1670-pixel-per-line resolution in luminance. The detail refresh rate would be as slow as five frames per second for some components but can be faster for others. The approximate proportions of the information in the detail channel would be: luminance, 75%; R-Y, 20%; and B-Y, 5%.

Summary

The psychophysical basic data has been gathered and a preliminary feasibility demonstration has been completed to support the concept of a fully compatible HDTV analog transmission system which can achieve a bandwidth reduction of about three or four to one. This system would allow the recording of HDTV images on standard high-band video tape recorders and allows compatible transmission of television with full 35 mm film resolution within a bandwidth available in two broadcast channels.

Even though the processing required uses a frame store in the HDTV receiver it is expected that within the time necessary to develop the HDTV system for the consumer, the cost of these stores will become reasonable. The use of this processing can provide completely compatible HDTV transmission in about one-quarter the spectrum space that would otherwise be required.

References

1. Breitmeyer, Bruno G. and Leo Ganz, "Implications of Sustained and Transient Channels for Theories of Visual Pattern Masking, Saccadic Suppression, and Information Processing." Psychological Review, 83, No. 1, 1976, pp. 1-36.

2. Fujio, Takashi and Keiichi Kubota, "Transmission Primaries and Signal Forms." NHK Technical Monograph, No. 32, June 1982, pp. 27-33.

A SINGLE RF CHANNEL STEREO REMOTE PICK-UP TRANSMITTER

Bradley Dick

KANU/KFKU

Lawrence, Kansas

One of the many problems facing any FM broadcaster today is the inability to easily transmit stereophonic programming from a remote location back to the studio. Although attempts have been made to create a stereophonic transmission system for Remote Pickup (RPU) uses, to date none have been completely successful. This paper will describe a new transmission system which will meet the needs of the FM (and now stereo AM) broadcaster when stereo transmission is needed from a remote location. One primary advantage of this system is its ability to use a repeater system and still only use two frequencies.

INTRODUCTION

For years broadcasters have had to rely on telephone lines for the transmission of program material from the remote broadcast site back to the studio. These telephone lines were often of poor quality, expensive, and required long lead times for installation. Additionally, there were always some locations where telephone service could not be provided. In those instances, the radio station either had to rely on 2-way radio equipment to relay the information back to the studio, or forget the broadcast. When the 2-way system was used, no amount of post production could improve the quality. It "sounded" like a 2-way radio transmission.

Recent years saw wider band, higher quality transmitters and receivers used in place of the 2-way equipment. The newer remote pickup equipment provided improved signal to noise ratios and better frequency response. The term RPU soon evolved into meaning a higher quality type of transmission system than that available from the 2-way equipment. Stations used the RPU equipment in any number of applications and in general were satisfied with its use. In all but a few special cases however, the available equipment limited the audio frequency response to 10 kHz, or less. Additionally, the audio performance of the equipment left much to be desired. The audio control sections were usually adequate only for voice. Music, especially anything other than "popular music" suffered from the simplistic audio control sections in the transmitters.

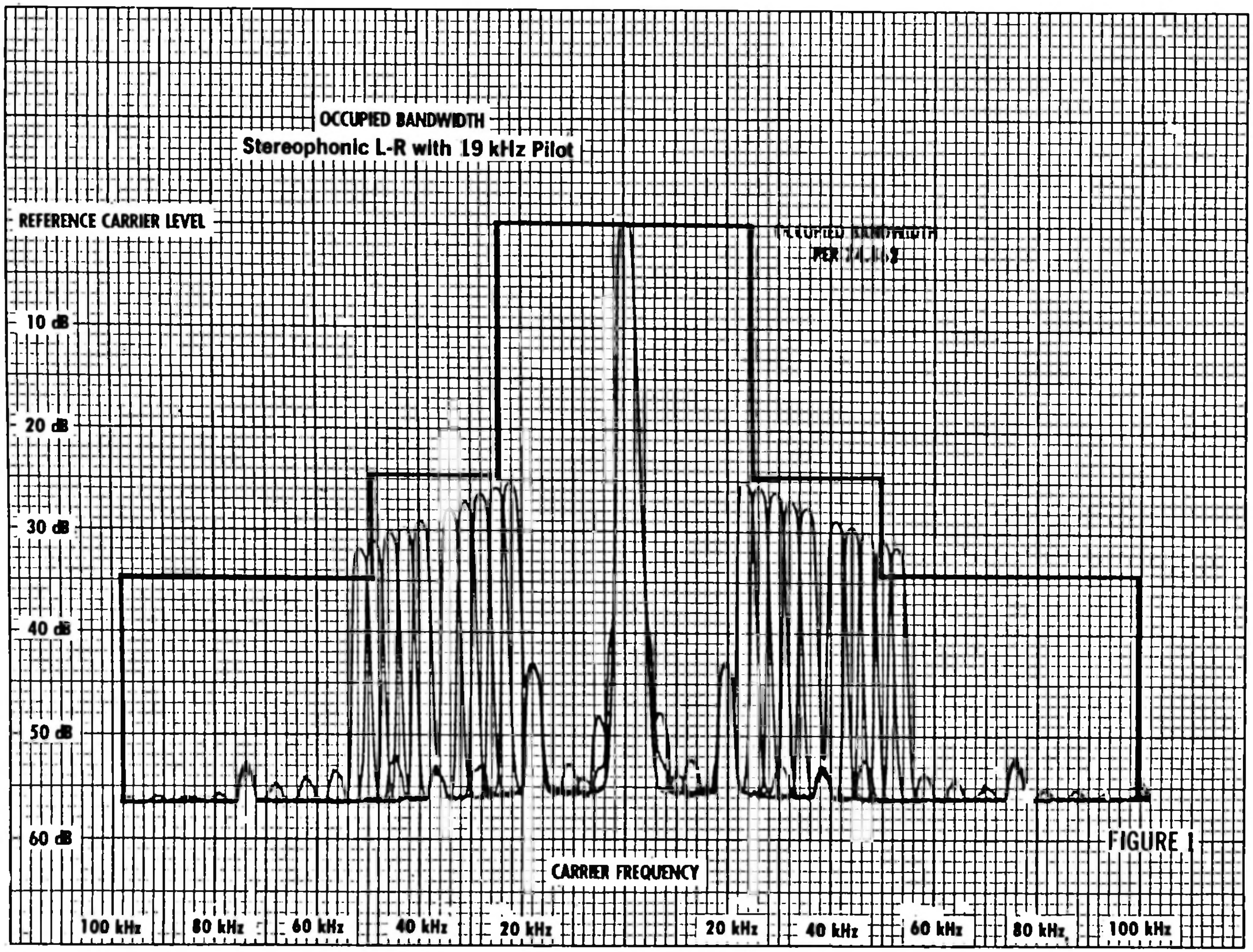
In addition, there remained one other aspect of remote pickup broadcasting which went unanswered for the FM broadcaster - stereophonic transmission. The FCC allocated two wide band channels, 450.925 MHz and 455.925 MHz, for the transmission of two wide band audio channels, and some broadcasters used them for stereo transmission. The system worked for a station in an area with little competition for spectrum space. For a station in a major market however, it was almost impossible to obtain the use of both of the wide band channels at the same time for a stereo broadcast. Additionally, if the station needed a repeater to get the signal back to the studio, then the whole prospect of stereo programming became impossible to realize.

CURRENT LIMITATIONS

The limitations imposed by Part 74 on even the so called wide band channels are quite severe. To transmit any stereo information within these limitations would be practically impossible. I say practically, because careful engineering <u>can</u> surmount these limitations.

Figure 1 shows an actual stereo composite signal transmitted at 450.925 MHz. This signal consists of a L - R audio signal imposed on a typical RPU transmitter. The S Channel limitations of Part 74 are also shown on this graph. By observing the outer skirts of the signal, it is obvious that the required bandwidth of the stereophonic 19 kHz signal exceed those limitations.

Investigations into the operating characterists of RPU transmitters shows a tendency on the part of the manufactures to remain overly cautious interms of using the permitted occupied bandwidth. In one case the transmitter, while being modulated at a level the manufacture called 100%


OCCUPIED BANDWIDTH
Stereophonic L-R with 19 kHz Pilot
REFERENCE CARRIER LEVEL
10 dB
20 dB
30 dB
40 dB
50 dB
60 dB
CARRIER FREQUENCY
100 kHz
80 kHz
60 kHz
40 kHz
20 kHz
20 kHz
40 kHz
60 kHz
80 kHz
100 kHz
FIGURE 1

modulation, was using only about 50% of the occupied bandwidth allowed by Part 74. This practice gave hope to the idea that it might be possible with modern technology to more fully utilize the available bandwidth and obtain a quality stereo signal.

STEREO TRANSMITTER DEVELOPMENT

The above problems caused KANU Engineering Services to begin investigating the possibility of developing a stereophonic high quality Remote pickup transmitter. The criteria for the new type of transmitter included:

1) System must operate in stereo on one S channel
2) System must provide ability for repeater use
3) System must provide a high quality signal
4) System must be cost effective
5) System must be licensable

After reviewing the available literature on remote pick-up equipment, it quickly became obvious that no manufacture had the necessary equipment to meet the above requirements. It was also discovered at the 1981 NAB Convention, that none of the RPU equipment manufactures were interested in developing such a system. In every case, they refused to cooperate in the development of such equipment. Every manufacturer stated that "It's impossible" to operate a stereo remote pickup system under the rules of the FCC.

It was obvious that given sufficient bandwidth, a composite stereophonic signal could be transmitted in the 450 MHz band. However, discussions with the FCC revealed that under no circumstances would they consider a waiver in order to provide the additional bandwidth which was needed for a standard composite stereo signal. With the current limitations in bandwidth, it was therefore obvious that some new method of stereo transmission was needed.

STEREO SIGNAL

At this point in the development, it was apparent that either we had to petition the FCC to provide waiver (to which they had already indicated an infavorable attitude) or come up with another system. Tests showed that if we were willing to trade off some small amount of upper frequency response, it might be possible to legally transmit a composite stereo signal within the limitations of Part 74.

The stereo transmitter was developed by using an Optimod 8000 as the first building block. First, the pilot frequency was moved from 19 kHz to 15 kHz. It was obvious from the beginning of the design process, that the audio had to be band limited if the transmitter was to be able to operate successfully within the requirements of Part 74. Since a 15 kHz pilot would be used to generate the stereo information, we decided to use a 10 kHz audio filter to protect the pilot from interference.

Originally, the Optimod was modified to operate with 10 kHz filtering provided by its own input circuits. While this method seemed to work, better dynamic audio performance could be obtained by using actual 10 kHz low pass filters. Inorder to obtain the kind of low pass filtering desired, several designs were examined.

An active filter could have been used and several designs were tested on the bench. However, the performance was insufficient to allow full frequency audio and still protect the pilot. In looking at the other options, it became apparent that a high quality low pass filter was needed. To provide the necessary performance characteristics, it would of necessity be quite complex. It needed to be very flat out to 10 kHz, yet provide a very rapid attenuation above 10 kHz. Finally, it needed to be small and passive.

A small company in Raytown Missouri, Filtronetics, specializes in custom passive filters. Consulations with their design department revealed that they could indeed build a filter to meet our requirements. The result of their work in shown in Figure 2. The response of the filter is within .38 dB at 10 kHz and yet is down 66.86 dB at 15 kHz. This filter provided the kind of control needed to maximize the audio performance of the transmitter. This filter was then installed on each input channel of the Optimod.

The other Optimod functions were left intact. Experience showed the system performed best when the audio control sections of the Optimod were used very little. The advantage of using the Optimod audio processing was that it provided a certain amount of gentle audio processing while also providing the very necessary function of safety limiting in the output stage.

Once the stereo signal was generated in the Optimod, it was still necessary to modulate a transmitter and develop the 450 MHz signal. Using a standard remote pickup transmitter, tests were conducted to determine what type of quality could now be obtained with the equipment. It was obvious from the very first test that problems still existed. While a true stereo signal was being transmitted through the system, serious degradation of that signal was taking place.

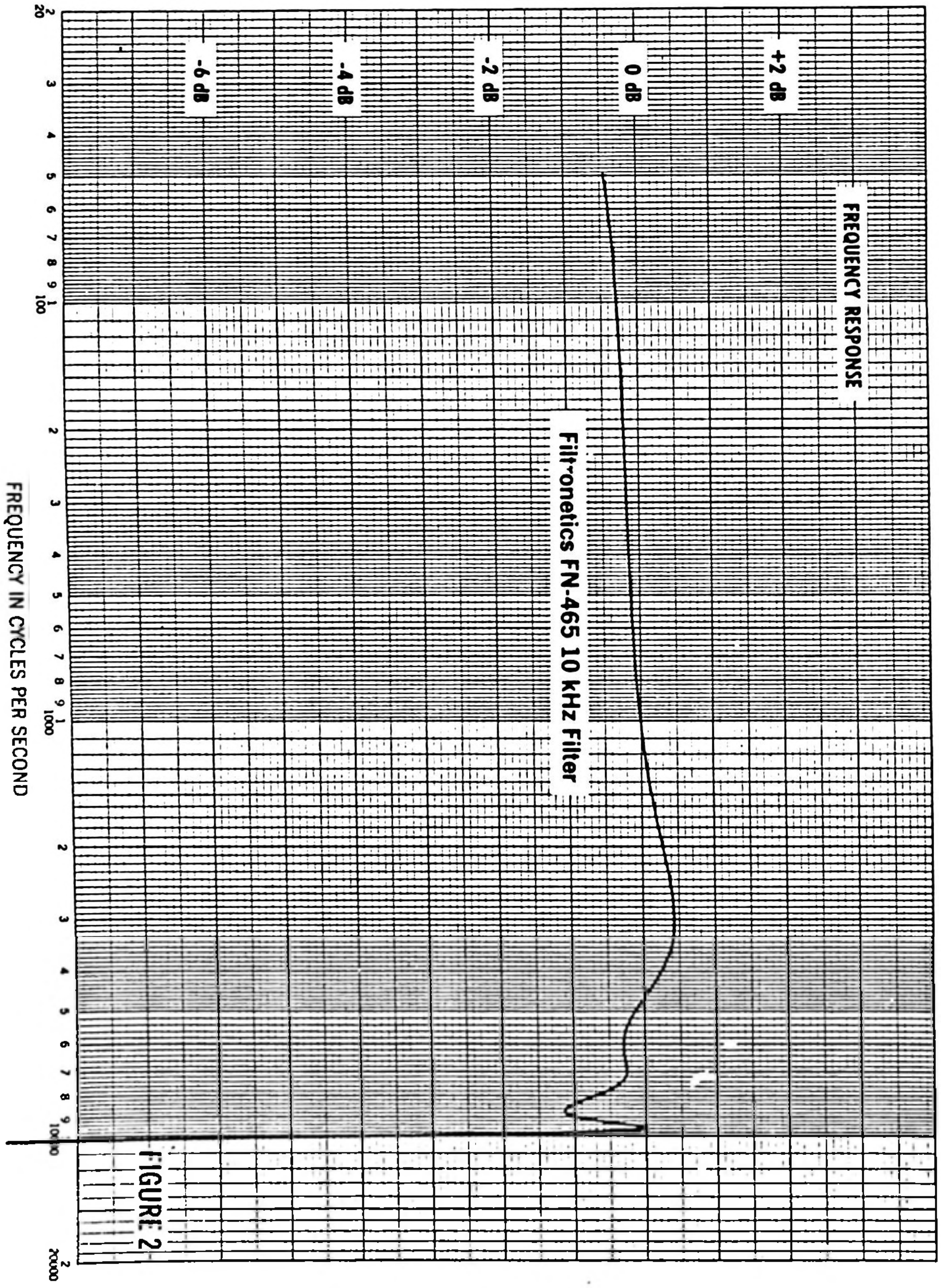

FREQUENCY RESPONSE
+2 dB
0 dB
-2 dB
-4 dB
-6 dB
Filtronetics FN-465 10 kHz Filter
FREQUENCY IN CYCLES PER SECOND
20
100
1000
10000
20000

FIGURE 2

The monophonic signal was of good quality, but the stereo portion of the signal was quite bad. Some form of interference appeared to be causing breakup of the L - R sidebands during transmission.

The lower trace in Figure 3 shows the frequency response of the original transmitter crystal. Notice the spurious response of the crystal at 10, 17 and 36 kHz. These spurs were the cause of the breakup of the L - R sidebands.

After discussing the problem with several people, we decided to see if a crystal manufacture would be willing to work with us to develop a crystal which would not have such problems. Working with KANU engineer Bob Pearson, American Crystal in Kansas City was able to make a crystal which met the tolerance requirements of Part 74, while at the same time, not have spurs within the critical area of the stereo baseband. The upper trace in Figure 3 shows the results of that work. Note the spurs have now been moved up in frequency, to just outside of the critical area of the stereophonic baseband.

RECEIVER IF FILTERS

Once the crystal problems were solved, it was possible to modulate the transmitter with an interference free composite stereo signal. Still, some receiver work was necessary. Prior to this point in the design process, a custom stereo decoder-driver circuit was built by Pearson. The new decoder-driver was necessary as no 15 kHz receivers were available.

The IF bandwidth of the receiver as supplied by the manufacture was insufficient. Tests showed poor response in the critical areas of the stereo baseband. Figure 4 shows the frequency response of the 85 kHz IF filter as it was received from the manufacture. A new 200 kHz IF filter was installed in place of the original unit and receiver performance was considerably improved. The modified receiver IF frequency response is shown in Figure 5. The total system audio frequency response is shown in Figure 6. This graph represents the total system performance of the right channel audio chain from transmitter input to receiver output.

At this point, the system was capable of transmitting a true working stereophonic signal. It was still not a usable stereo signal however, because of the poor signal to noise ratio. The measured signal to noise was approximately 32 dB which would be unacceptable for a monophonic signal, let alone a stereophonic one. To obtain the necessary signal to noise ratio, it was again necessary to deviate the RF

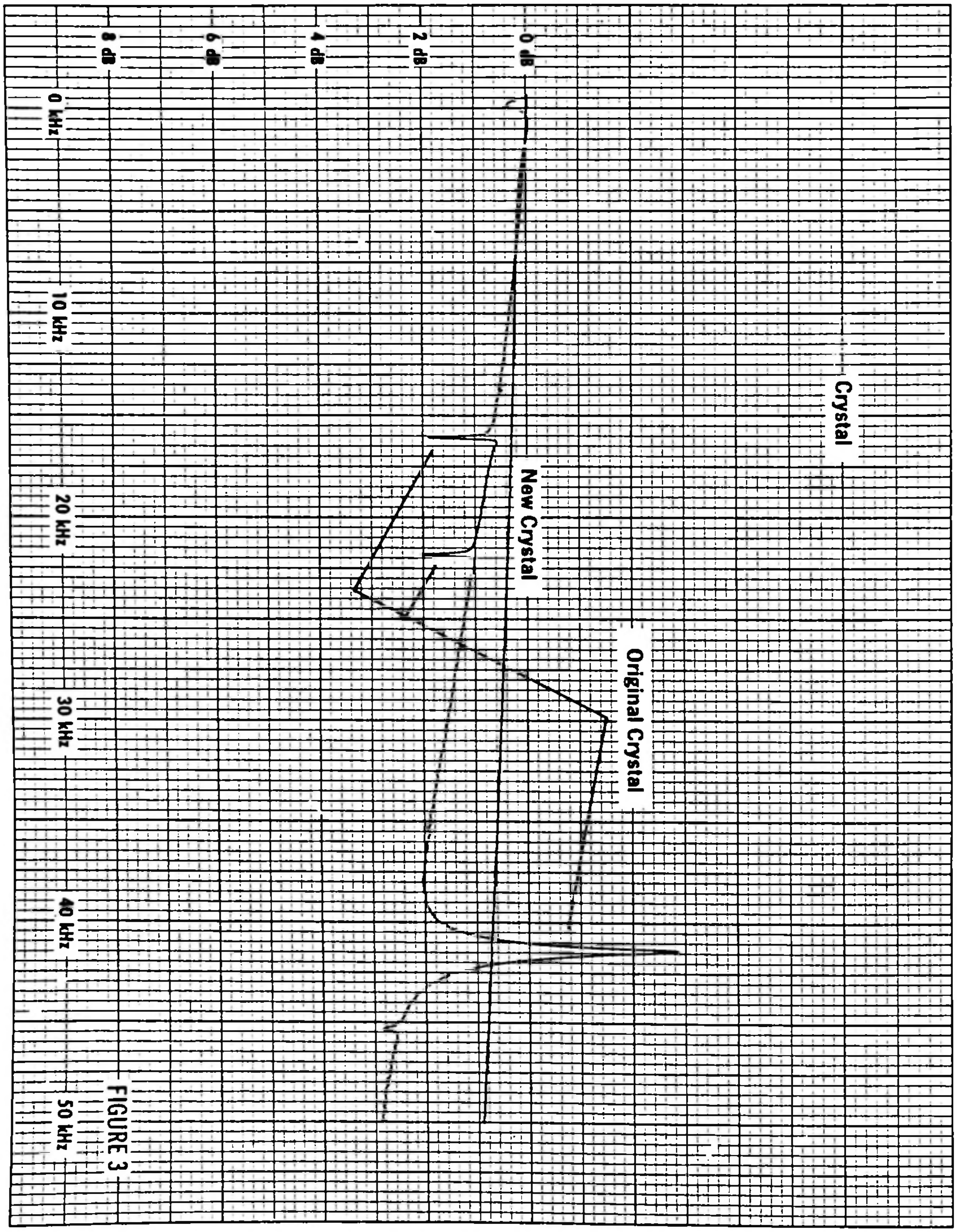
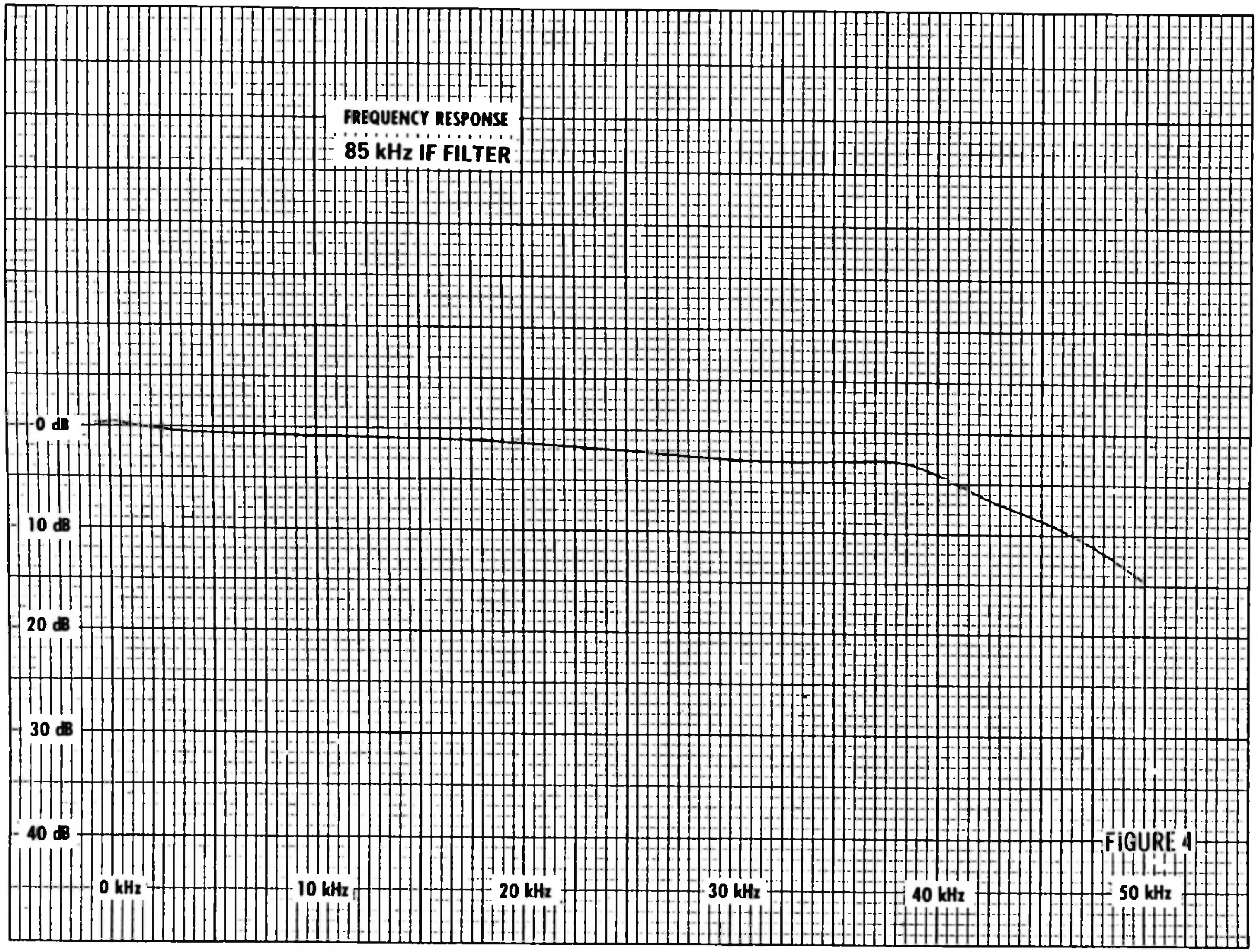
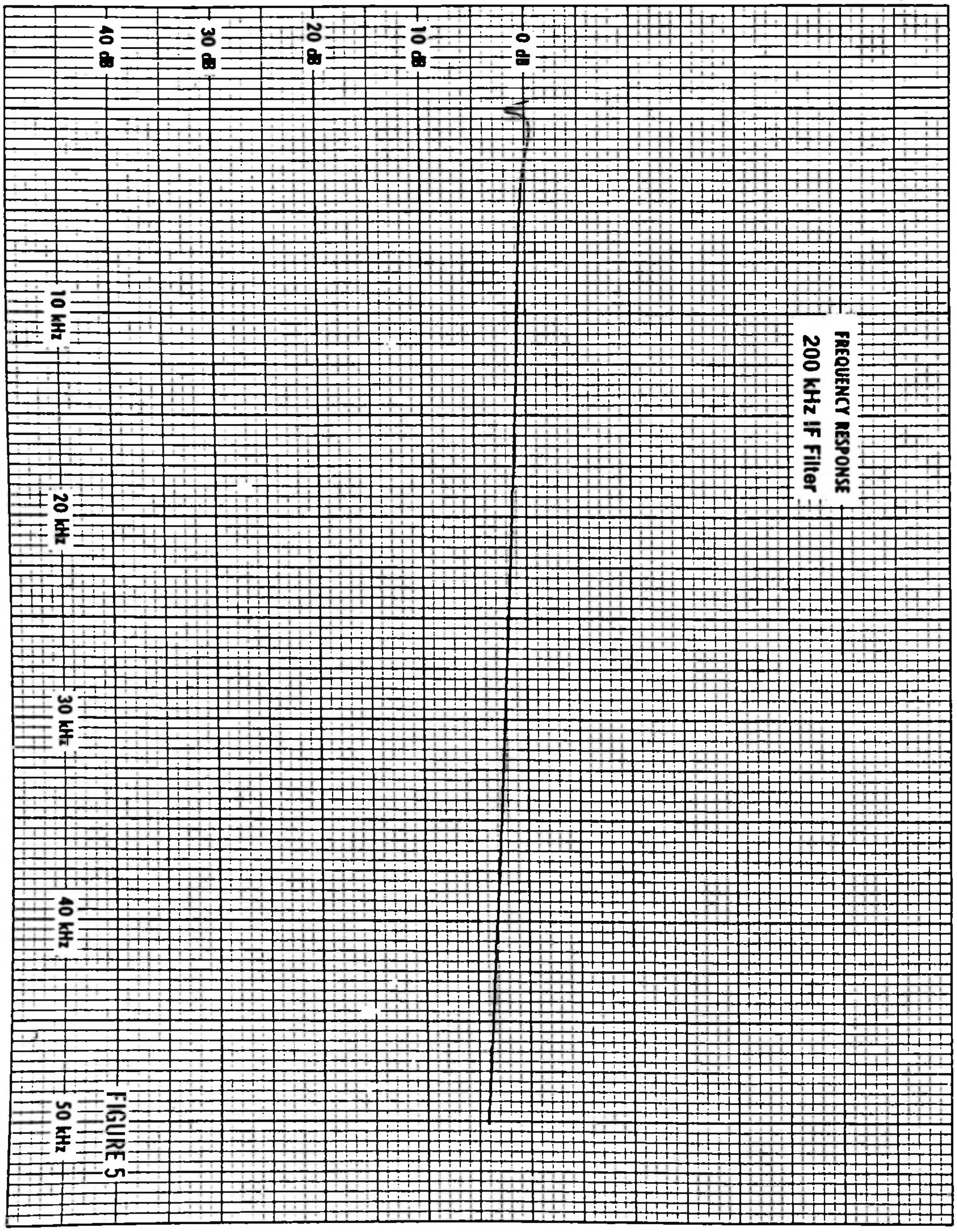
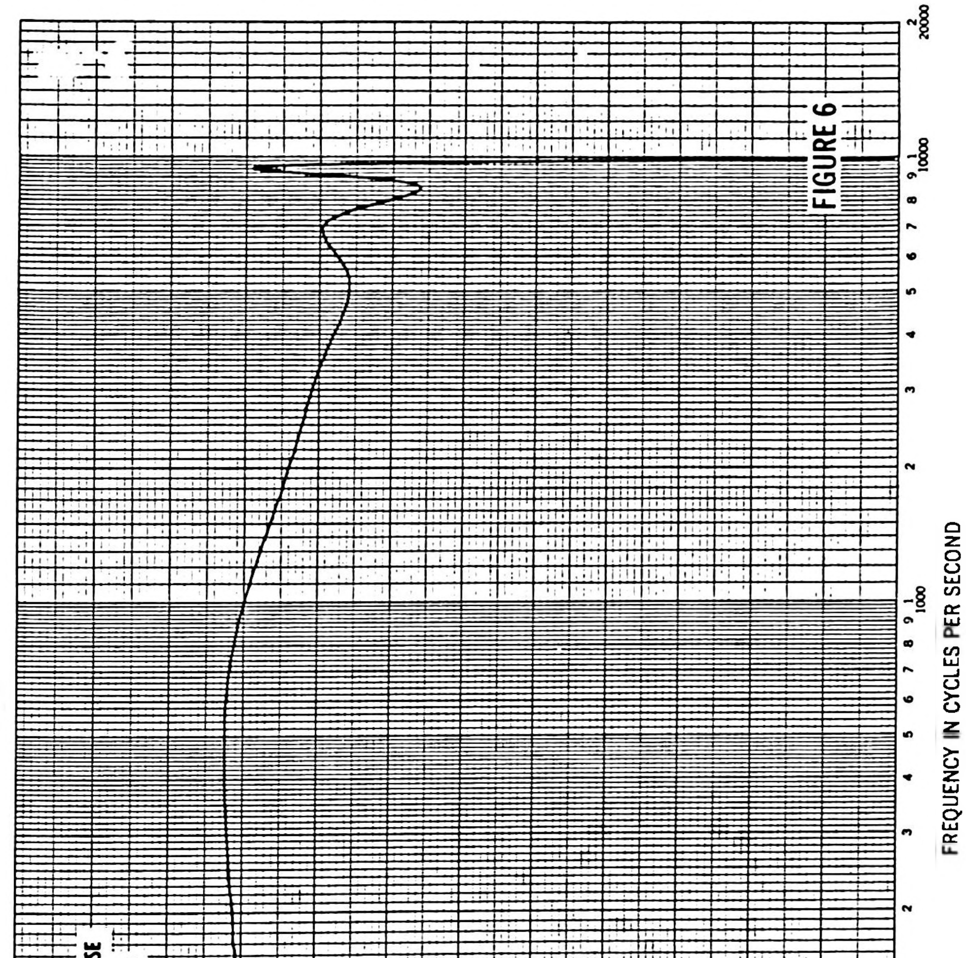

0 dB
2 dB
4 dB
6 dB
8 dB
0 kHz
10 kHz
20 kHz
30 kHz
40 kHz
50 kHz
Crystal
New Crystal
Original Crystal

FIGURE 3


FREQUENCY RESPONSE
85 kHz IF FILTER
0 dB
10 dB
20 dB
30 dB
40 dB
0 kHz
10 kHz
20 kHz
30 kHz
40 kHz
50 kHz
FIGURE 4


FREQUENCY RESPONSE
200 kHz IF Filter
0 dB
10 dB
20 dB
30 dB
40 dB
10 kHz
20 kHz
30 kHz
40 kHz
50 kHz
FIGURE 5



FREQUENCY RESPONSE
Total System
FIGURE 6
+2 dB
0 dB
-2 dB
-4 dB
-6 dB
20
100
1000
10000
20000
FREQUENCY IN CYCLES PER SECOND

carrier in excess of the limitations imposed by Part 74. When the deviation was reduced to the point where the limitations of Part 74 were met, then the signal to noise was unacceptable.

SIGNAL TO NOISE IMPROVEMENT

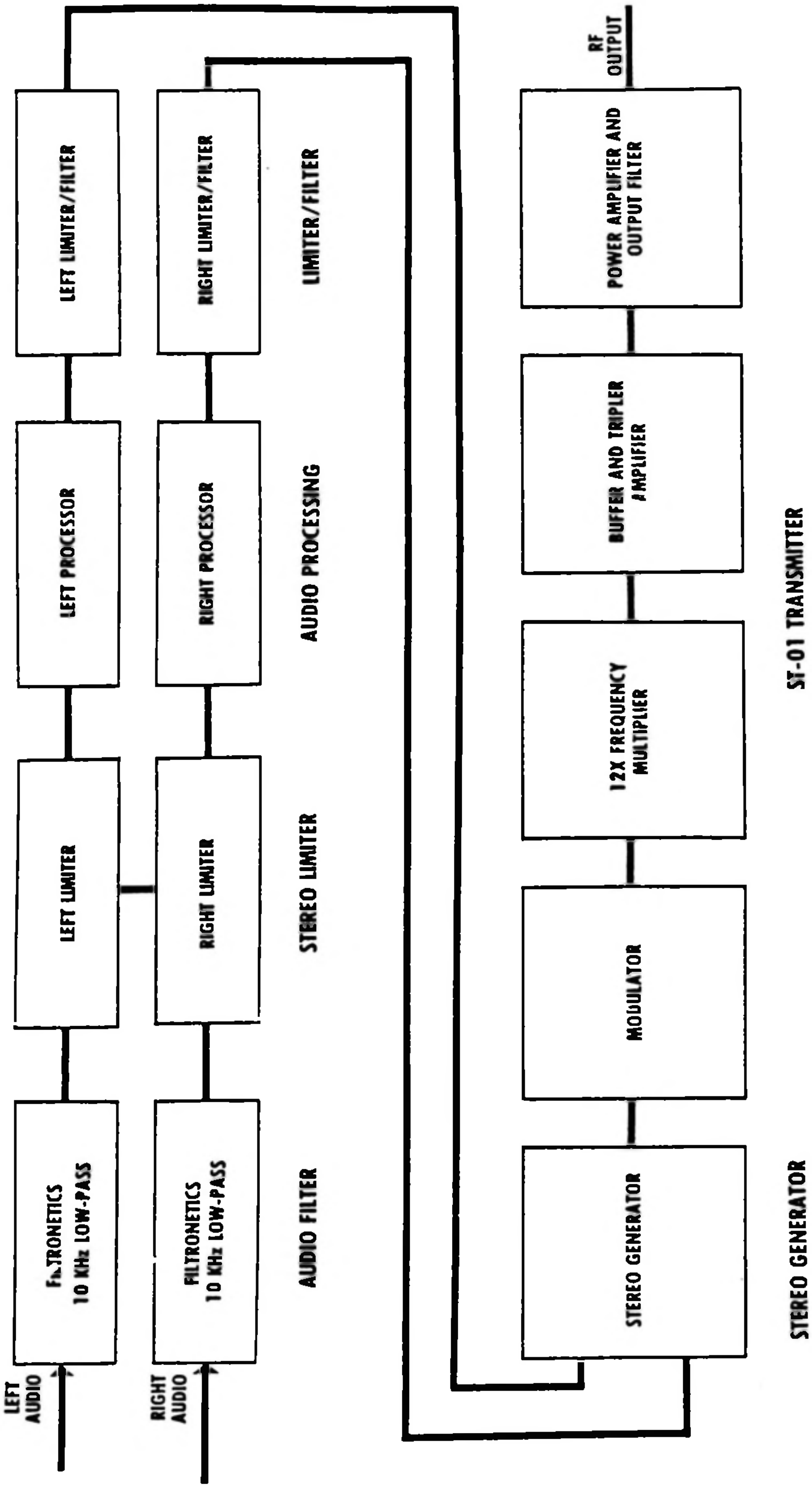
At this stage of development we had a stereophonic transmitter capable of generating a true stereo signal at 450 MHz. And most importantly, the signal conformed to the limitations of Part 74. However some form of improvement was still needed in the signal to noise ratio.

KANU then undertook the development of an audio processing system which could sufficiently control the audio so that a wide dynamic range signal could be transmitted and still not be lost in the noise. Figure 7 shows the block diagram of the final transmitter. The block labeled Audio Processing indicates the place in the audio chain where special audio processing takes place before it is applied to the transmitter. This processing system is currently under going patent proceedings, and cannot be fully described here. The special circuits do however, provide the necessary processing to allow the final system to provide a signal to noise ratio in excess of 80 dB.

EQUIPMENT SUMMARY

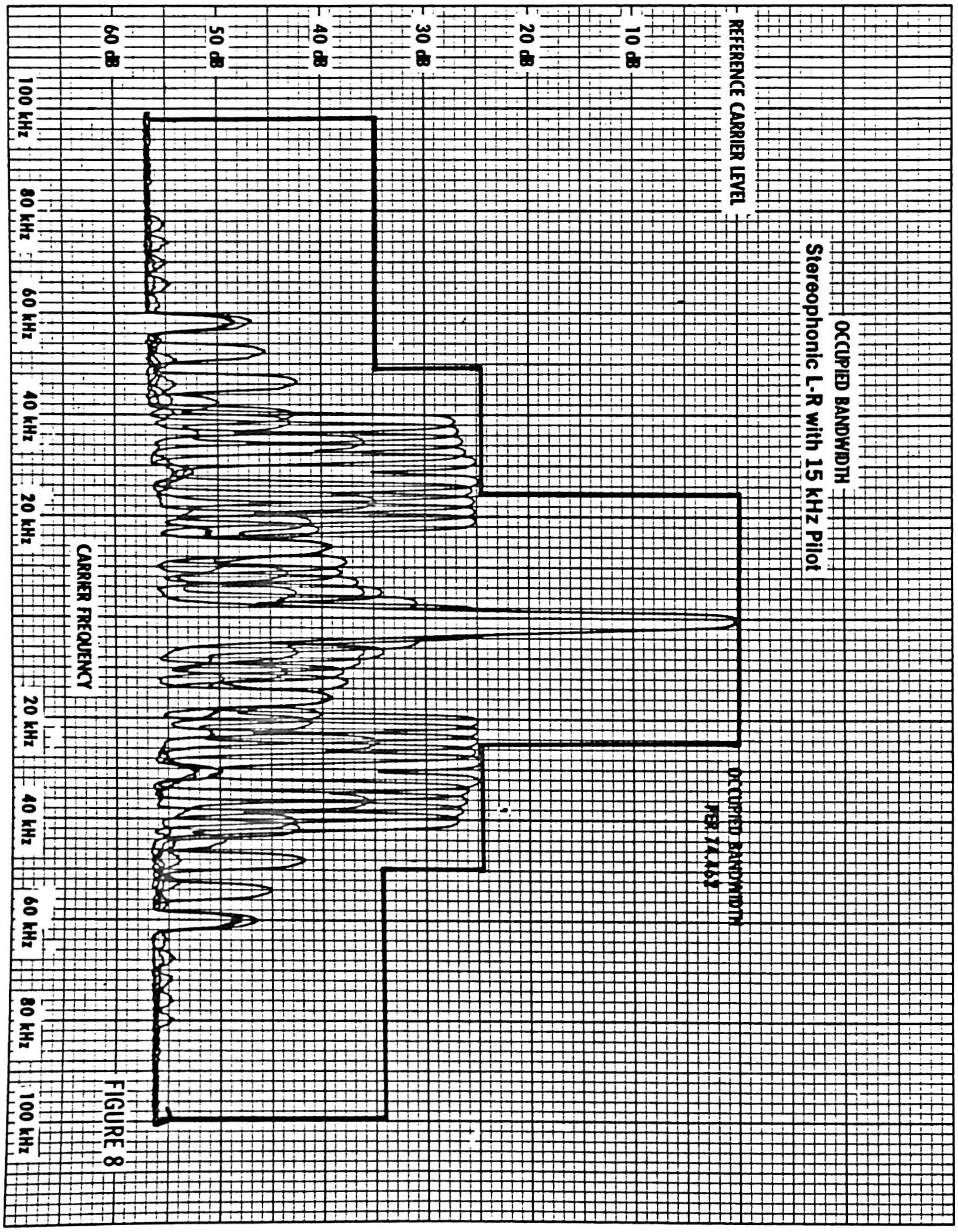
To confirm the critical parameters of the new transmitter, tests were conducted to insure that the limitations of Part 74 would not be exceeded. Figure 8 shows the occupied bandwidth of the transmitter when being modulated with a L - R signal (the most critical test). Notice the restrictions of Part 74 are clearly met. Figure 9 shows full modulation with a right channel only signal being transmitted. Figure 10 shows the transmitter being fully modulated with a L + R signal. Again, in all cases the transmitter fully meets the restrictions on occupied bandwidth as defined by Part 74.

The criteria of the original design had now been met. The transmitter was capable of providing a wide band stereophonic signal on a single RF carrier. The requirement of only one RF carrier insured that a repeater system could be used which would vastly increase the operating range of the system. The system allowed KANU to license the system with the FCC and still meet all Part 74 requirements. The trade off made in frequency response was judged to be acceptable because of the increased programming opportunities. After being in use for many months now, nothing but excellent reports have been received.



BLOCK DIAGRAM OF STEREO RPU TRANSMITTING SYSTEM

FIGURE 7


OCCUPIED BANDWIDTH
Stereophonic L-R with 15 kHz Pilot
REFERENCE CARRIER LEVEL
10 dB
20 dB
30 dB
40 dB
50 dB
60 dB
100 kHz
80 kHz
60 kHz
40 kHz
20 kHz
CARRIER FREQUENCY
20 kHz
40 kHz
60 kHz
80 kHz
100 kHz
OCCUPIED BANDWIDTH
PER 74.463
FIGURE 8

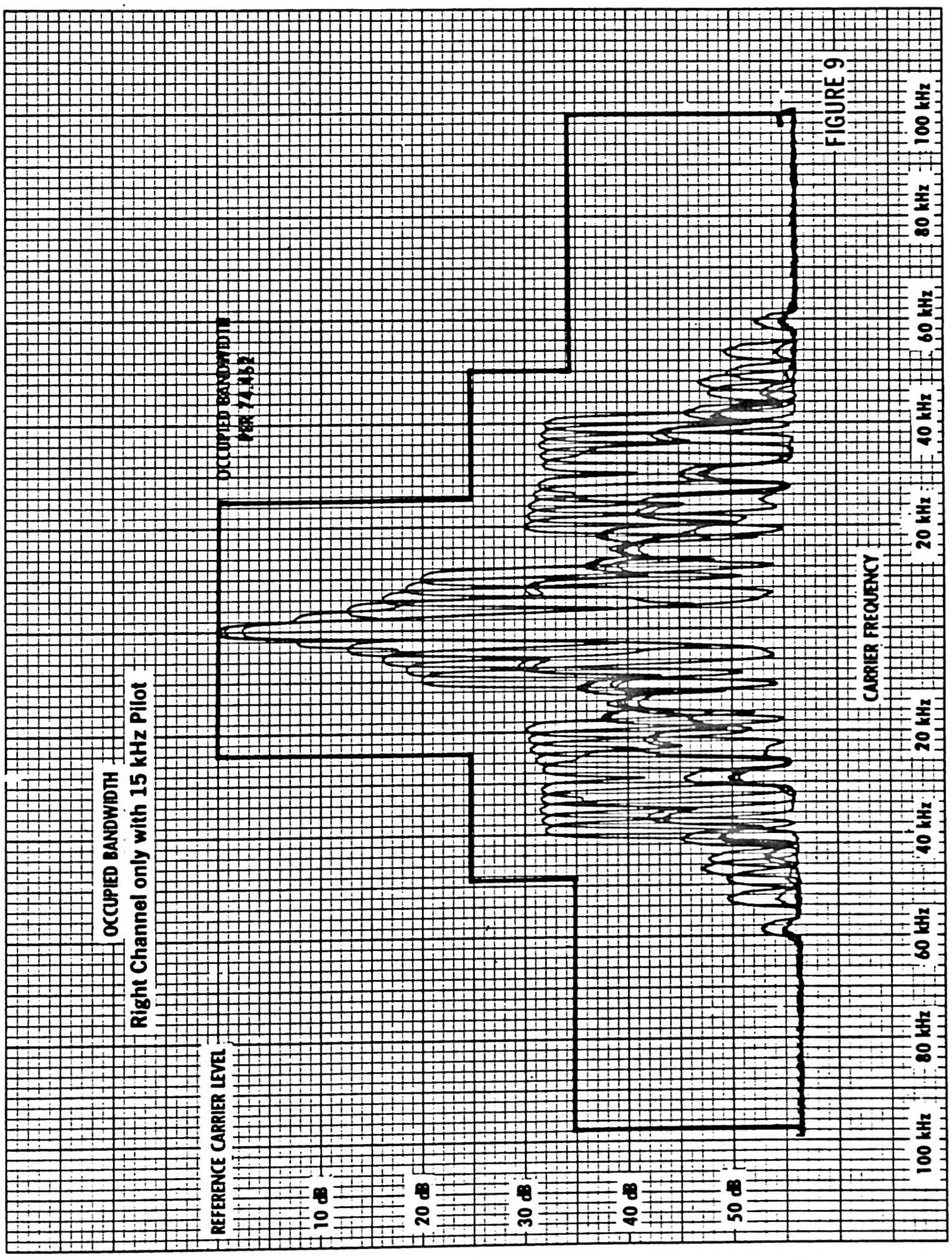

OCCUPIED BANDWIDTH
Right Channel only with 15 kHz Pilot
REFERENCE CARRIER LEVEL
10 dB
20 dB
30 dB
40 dB
50 dB
100 kHz
80 kHz
60 kHz
40 kHz
20 kHz
CARRIER FREQUENCY
20 kHz
40 kHz
60 kHz
80 kHz
100 kHz
OCCUPIED BANDWIDTH
PER 73.44

FIGURE 9

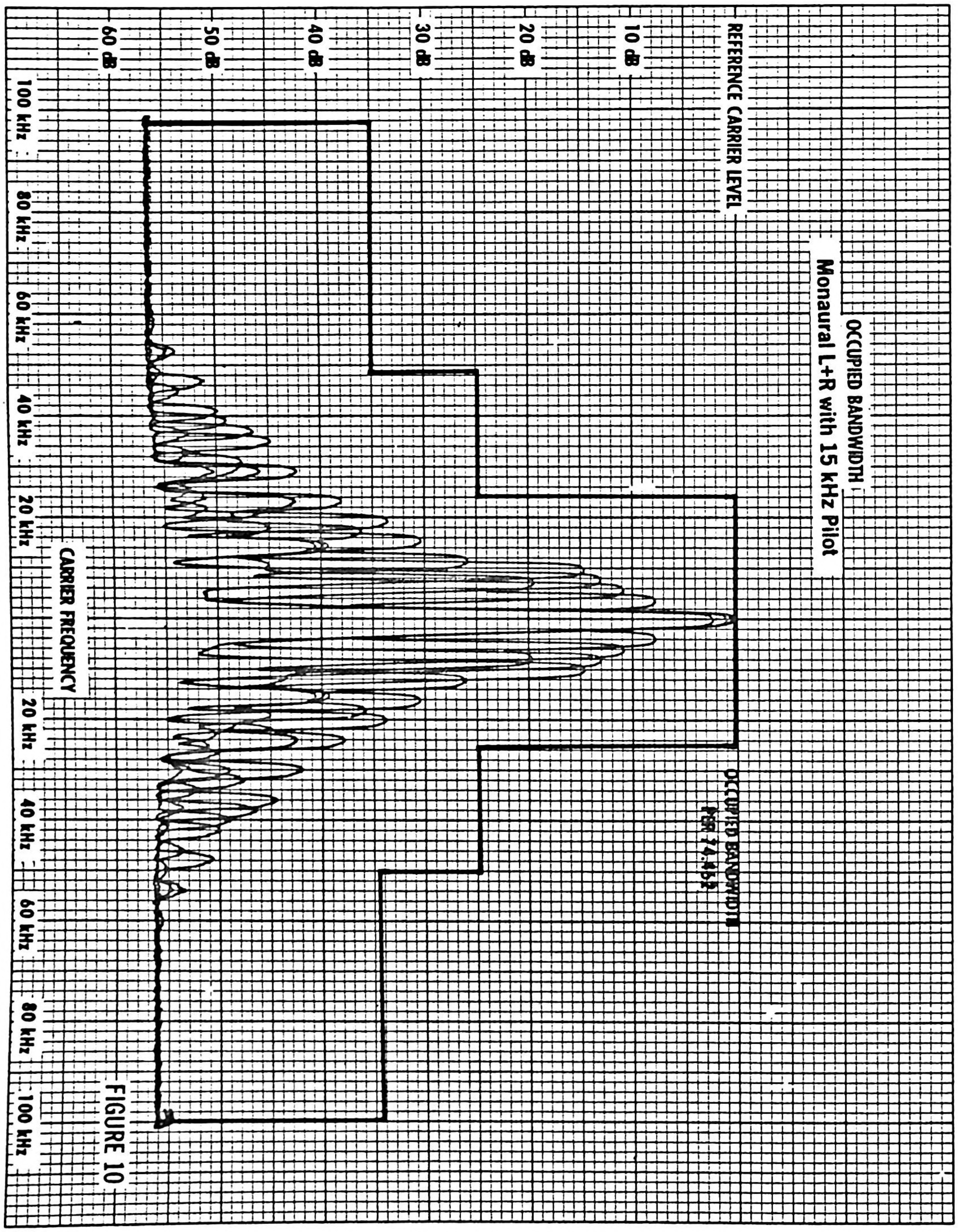

OCCUPIED BANDWIDTH
Monaural L+R with 15 kHz Pilot
REFERENCE CARRIER LEVEL
10 dB
20 dB
30 dB
40 dB
50 dB
60 dB
100 kHz
80 kHz
60 kHz
40 kHz
20 kHz
CARRIER FREQUENCY
20 kHz
40 kHz
60 kHz
80 kHz
100 kHz
OCCUPIED BANDWIDTH

FIGURE 10

The system is used to regularly broadcast stereophonic jazz and classical broadcasts from a variety of locations to the KANU audiences. Bluegrass broadcasts are undertaken on a regular basis from locations which would not otherwise be possible if telephone lines were required. Even sports broadcasts can now be done with ease. The broadcast of live basketball in stereo is an event which in not only quite lively, but can also be very sellable. The addition of the game ambience adds a great deal to the enjoyment of the game to a radio audience. The ability of a radio audience to follow the action up and down the court is (at least for KANU audiences) quite a treat.

OTHER OPTIONS

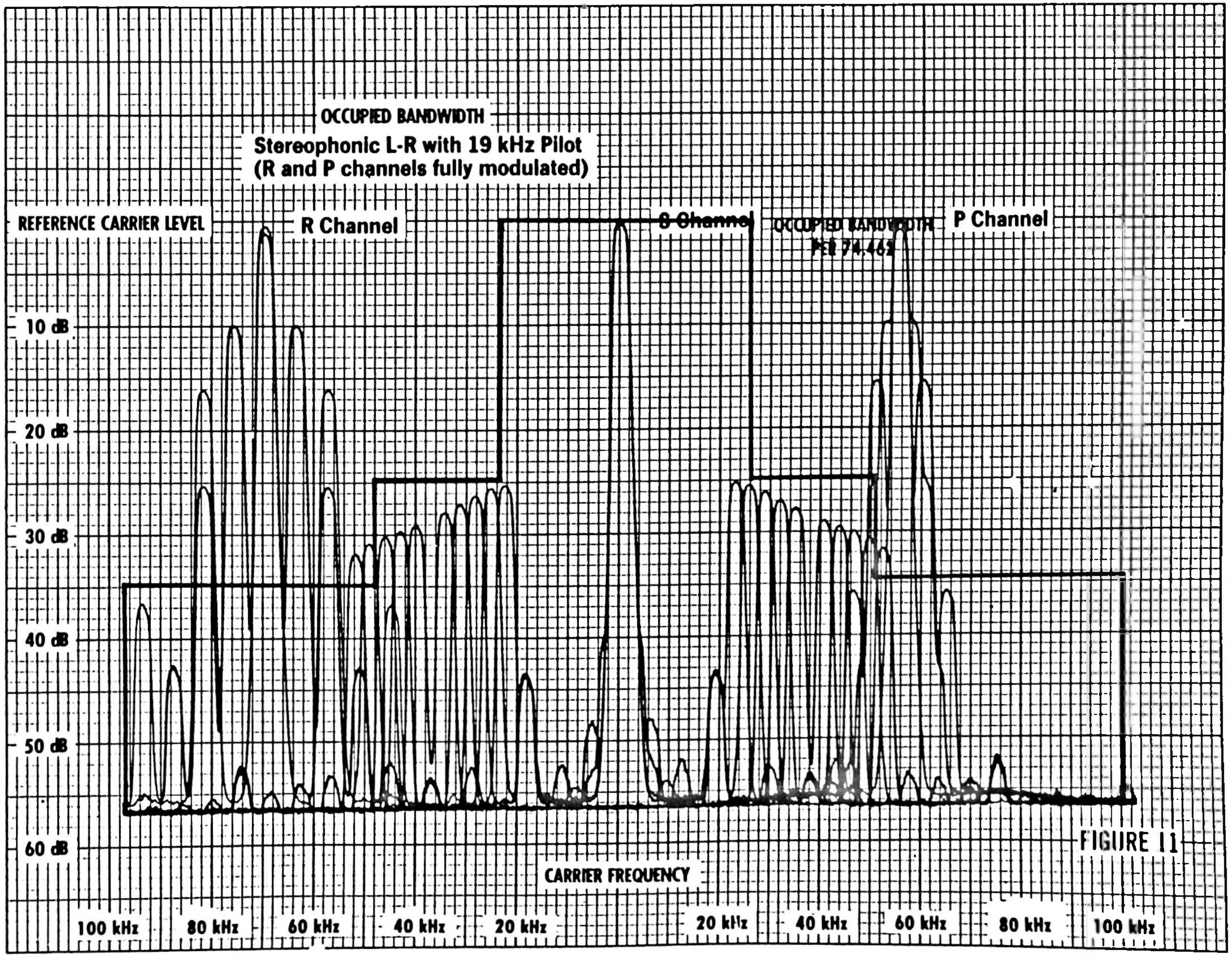
The numerous tests, and experimentation undertaken by KANU Engineering Services would not have been necessary had different bandwidth limitations been placed on the two S channels. The final section of this report will attempt to outline the severity of interference that might be encountered if different bandwidth limitations were possible. It will be shown that any interference which might be encountered is for the most part insignificant.

Figure 11 shows a 19 kHz L - R composite stereo signal with the limitations of Part 74, and the two adjacent R and P channels. Notice how the side bands extend just slightly beyond the occupied bandwidth restrictions of Part 74. These R and P channels have been fully modulated in accordance with Part 74. It is apparent that the stereophonic signal crosses into the bandwidth of both the R and P channels by a small amount. It is this amount of "interference" which needs investigation.

The critical aspect to consider here is the amount of interference which might be present. Certainly there is the possibility of having interference in any transmission system. The key to successful implementation of this system, is the very small amount of interference which would be caused to the adjacent channels.

Figure 12 shows a spectral display of four 450 MHz carriers. The center carrier is a 19 kHz composite L - R stereophonic signal modulated at a level compatible with the requirements of the KANU established guidelines. The adjacent two R and P channels are also shown.

Inspection of the mutual interference of the two lower R channels shows interference to be 35 dB below carrier level. In other words, the two R channels are interfering with each other at a point 35 dB below the carrier level.


OCCUPIED BANDWIDTH
Stereophonic L-R with 19 kHz Pilot
(R and P channels fully modulated)
REFERENCE CARRIER LEVEL
R Channel
S Channel
OCCUPIED BANDWIDTH
PER 74.462
P Channel
10 dB
20 dB
30 dB
40 dB
50 dB
60 dB
CARRIER FREQUENCY
100 kHz
80 kHz
60 kHz
40 kHz
20 kHz
20 kHz
40 kHz
60 kHz
80 kHz
100 kHz
FIGURE 11

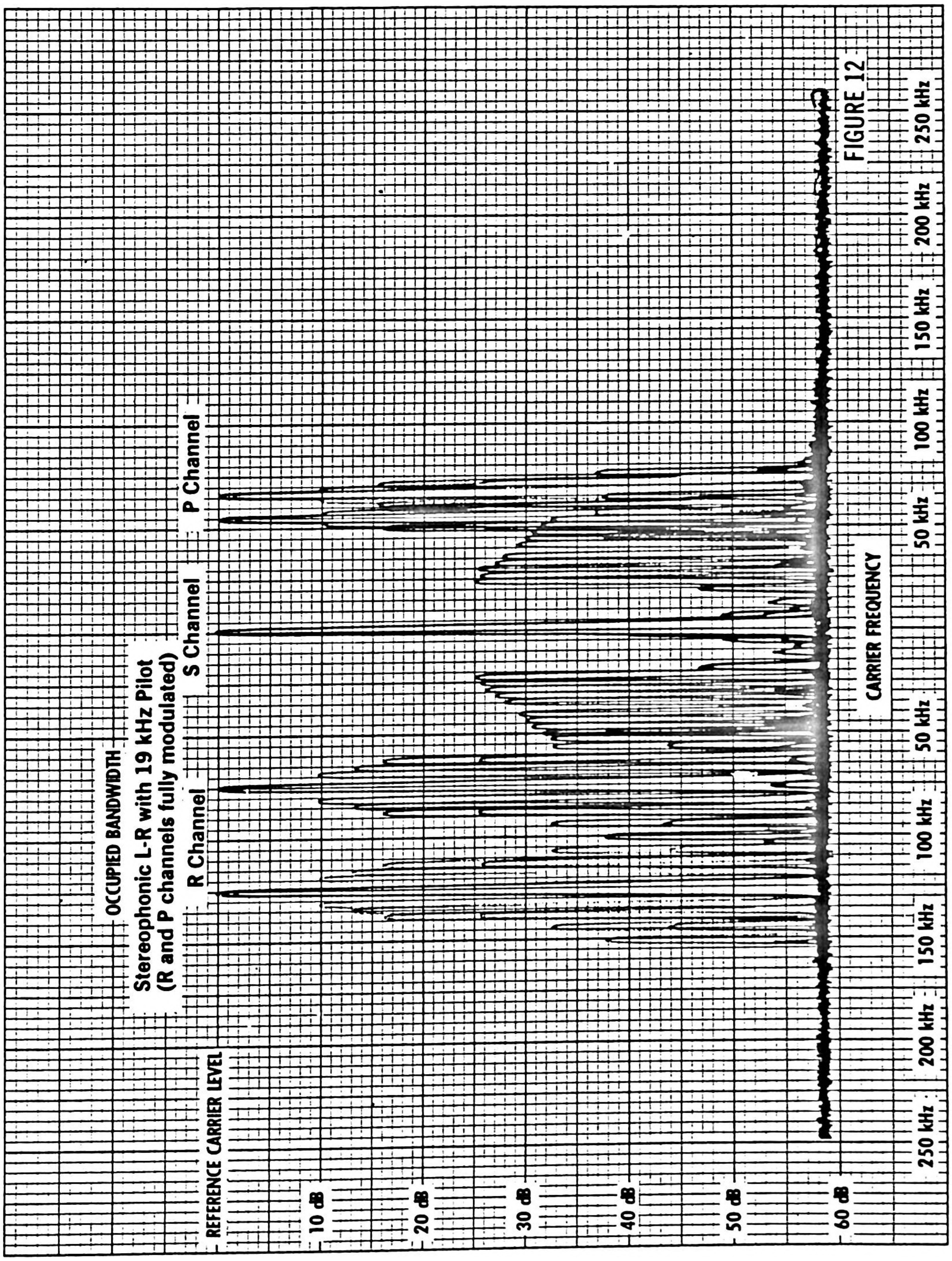

OCCUPIED BANDWIDTH
Stereophonic L-R with 19 kHz Pilot
(R and P channels fully modulated)
R Channel
S Channel
P Channel
REFERENCE CARRIER LEVEL
10 dB
20 dB
30 dB
40 dB
50 dB
60 dB
250 kHz
200 kHz
150 kHz
100 kHz
50 kHz
CARRIER FREQUENCY
50 kHz
100 kHz
150 kHz
200 kHz
250 kHz

FIGURE 12

Inspection of the same problem with the upper P channels reveals a similar condition. The two P channels are causing mutual interference at a level of only 35 dB below the carrier level. Since these two channels are for tone signaling purposes, this interference does not seem to be a problem. Close resolution on this graph is difficult because of the limitations on the chart recorder.

A close inspection of the interference of the stereo signal into the two adjacent channels shows a very interesting phenomena. The interference caused to the lower R channel by the stereo signal is so low that it would not likely be discovered by anyone operating on that R channel. Plus, the limitations of the receiver IF filtering would likely restrict any audible signals to a level far below what would be detectable. The exact same occurrence takes place with the P channel. The interference level from the S channel to the P channel is over 30 dB below the carrier level. Since the P channel is only for tone modulation, the interference would likely go undetected.

There is an other option which should be considered. Tests conducted near Kansas City and Topeka Kansas showed no use of the P channels. The channels were monitored for a period of several weeks and not one occurrence of traffic was detected over that time. While this does not necessary validate the lack of use of the channels, it is important to look into ways to maximize the use of the available spectrum. In this case all occupied bandwidth parameters for a 19 kHz stereophonic transmission system could be met if two small changes in Part 74 were permitted.

First the carrier frequency of the S channel would be moved up 4 kHz. Second, the first adjacent P channel would be eliminated. By shifting the S channel slightly up in frequency, no interference in caused to the lower R channel and the single P channel. Again the advantages offered by this small change would far out weigh the loss of the single P channel. This option would be much more spectrum efficient than the current method of requiring a broadcaster to use two frequencies and 200 kHz in the space where half of that is really required.

The final option which might be considered would be a simple change to permit operation on a "non interference basis". Operation under this system of transmission would allow the use of the 19 kHz transmitter and the wider bandwidth audio. It would be undertaken on the basis that it would not cause interference with the adjacent R and P channels. If interference did occur, it would be simple matter to turn off the pilot and revert to monaural operation.

Considering the amount of operation within the 450 MHz bands, it is important to utilize the frequencies as effectively as possible. The current practice of requiring a station to use two transmitters, two receivers, and two operating frequencies is outdated and ineffective. The proper operation of a composite stereophonic transmitter on either of the 450 MHz S channel frequencies could permit as many as twice as many stations to originate remote broadcasts.

The current state of technology is such that improved utilization of the available spectrum space could be had by a simple waver or modification of the rules of Part 74. The changes requested could be undertaken with a minimum of effort and would offer to the broadcaster many advantages.

SUMMARY

The 15 kHz pilot system of stereophonic transmission disserves consideration as a viable method of operation. The system is simple, effective, inexpensive and versatile. providing definite advantages to any broadcaster interested in stereophonic programming. It allows him to operate in either stereo or monaural as the situation permits. And, it allows him to do so at a reasonable expense. Little changes are required with current transmitters and receivers and the equipment is already available.

Any broadcaster who is interested in stereophonic remote broadcasting should make his interests known to the NAB, SBE and the FCC. The Commission is unlikely to initiate any interest in spectrum usage changes without strong impetus from the industry.

CREDITS

A great deal of credit must go to the Engineering staff at KANU. Without the efforts of Bob Pearson and Jim Todd, this project would never have come to fruition.

A NEW TYPE OF INFORMATION DELIVERY NETWORK

Jack R. Taub

National Information Utilities

McLean, Virginia

INTRODUCTION

The "last mile" of any data communications network is often an expensive bottleneck. While long haul transmission has benefitted from a host of new technologies, the local leg still consists primarily of individual pairs of wires.

National Information Utilities (NIU), in partnership with National Public Radio (NPR), is developing a nation-wide network which will offer an extremely low cost alternative to the last mile problem, along with a unique set of communications features. Combining NPR's satellite distribution system and radio stations with NIU's FM subcarrier technology produces a network capable of simultaneous, point-to-multipoint communication to a virtually unlimited number of recipients.

The INC network will employ FM radio subcarriers (FM Subsidiary Communication Authorization's or FM SCA), with proprietary NIU hardware for local distribution, and satellite transmission for intercity links. This combination will create an addressable, one way distribution network ultimately capable of reaching any town or city in the US with an FM radio station. No other combination of cable, broadcast TV or direct broadcast satellite has this capability, or is expected to in the near future. However, NIU will have the capability to integrate other technologies into the network, as needed for maximum flexibility.

To receive data over the network, a customer would need only to install a small, inexpensive receiver-demodulator (receiver) which will interface with most CRTs, printers and computers. Because each receiver is individually addressable, data put into the system would be available only at desired receivers. Data transmission and error rates are equal or better than that of a dedicated (leased) telephone company line.

The ease and low cost of simultaneous multipoint transmission will be very attractive to any business sending identical information to many locations on a daily basis, e.g., news services, retail chains, banks, etc. In addition, the unique combination of all features of the network, i.e., addressability, hard copy, high reliability, broad coverage, and remote activation capability, will make the network valuable to other classes of users who may not be attracted solely by inexpensive multipoint distribution capability.

The functional components of the communications technology which will be utilized by the National Information Utilities Corporation through its subsidiary, INC Telecommunications, include the FM SCA subcarrier, receivers and controllers. These are based on a number of proprietary skills and technologies, some of which NIU expects to be patentable. The following sections describe these technologies.

SUBSIDIARY COMMUNICATION AUTHORIZATION

The use of FM SCA's for analog signals has existed since the early 1960's. However, it was not until 1975 that the FCC expanded the permissable use to include slow scan TV, teletype, facsimile, etc. - so called "visual services". Some of the original applications are foreign language broadcasts and reading services for the visually impaired. It has only been more recently that digital FM SCA has proven workable. Even now, high speed (over 2400 bits per second (bps)) transmission, is still essentially unavailable for commercial applications. A few organizations are currently operating small, low speed, networks for banks, supermarkets and the agricultural industry. NIU has had FM SCA on trial use at 4800 bps for 68 branches of the Suburban Trust Bank in Maryland for more than a year.

Operation

An FM SCA, or subcarrier, works by injecting a separate FM signal into a station's main carrier at the broadcast frequency plus 67kHz or 92Khz, which then can be received by users with a special tuner, which contains a demodulator to decode the SCA signal.

This additional signal must be of low enough power and bandwidth so as not to interfere with the main audio channel. For an analog sideband this is not a problem, but the requirements of a digital signal are just the opposite. Digitization inherently increases the required bandwidth, and this need grows as the speed of transmission increases, making high speed digital transmission technically difficult using an FM SCA.

NIU recently completed a series of reliability tests on its 4800 bps technology with excellent results - an average bit error rate of about 10 to the minus 6 (better than a leased conditioned telephone line).

These tests confirmed that high speed transmission using an FM SCA is possible, and paved the way to develop an even faster system. NIU is now working on a 9600 bps evolution of the original 4800 bps technology. In addition to nearly doubling the effective data rate, the new system will incorporate a powerful error detection scheme and proprietary protocol which will enable INC to create several classes of service tailored to the usage, speed and reliability requirements of different customers. For example, the system can offer a low speed 300 bps newswire as well as a highly reliable 9600 bps service for transfer of critical financial information. Both the high speed technology and protocol are expected to be patentable. The station control and required modulation equipment and software are under development by NIU and will be ready for production in the first quarter of 1983.

THE INC NETWORK

The creation of a local network is technically quite simple and requires only the installation of NIU equipment at the FM station and a supply of receivers (see figure 1). Users would feed messages to the local station via the phone system. An FM station will usually be able to transmit acceptable FM/SCA data over a 30 to 60 mile radius, depending on antenna height, station power, and local topography. Once a message reaches the receiver, it can be temporarily stored in buffer memory while it is being sent out at a speed compatible with the attached computer or peripheral. An individual FM station has the capacity to support many different types of users simultaneously with little or no increase in the time required to transmit a message.

Satellite System

A national system will be formed using NPR's satellite system. NPR has over 240 affiliated FM stations that are all potential local networks, and each of them is linked to NPR in Washington, D.C. via the Westar IV satellite and individual downlinks (see figures 2 and 3) . In addition, there are 18 regional uplinks that allow transmission "upstream" from a region back to NPR. Where an NPR station is not available, a commercial station would be used.

Network Characteristics

The INC data delivery system is designed to provide a new, nationwide, one-way delivery of a variety of data services. The system is constructed around one or more backbone data channels operating at 9600 bps each. This channel (or channels) is piggy-backed over the NPR satellite system and available to the over 240 affiliated NPR stations currently equipped with earth stations.

The modulation method used for the 9600 bps channels involves special spectrum-shaping techniques, designed so that no interference is produced when modulated on the SCA. For convenience, the same modulation method is used for the satellite portion of the system. This has the advantage that in some cases it will be possible to connect directly from the output from the satellite demodulator to the SCA modulator input without any further equipment in the station.

The 9600 bps data channel is assembled by a control computer connected to the NPR Network Control Center. This computer collects the data to be transmitted from as many sources as required, including:

* X25 data lines from TELENET/TYMNET
* Dial-up lines
* Leased lines

The data stream is assembled by the control computer and the output passed to a signal shaper to provide a modulating signal with the desired spectral characteristics. At the local station the signal is taken from the earth station and in the simplest case it is applied directly to the SCA modulation input. (It may be necessary to use some form of limited distance modem if the earth station and transmitter do not have a low distortion communication channel between them.)

In most cases the local station will wish to insert data for local delivery. In these cases it will be necessary to interpose a small NIU computer between the demodulator and the SCA. This computer is accessed by dial-up lines or a local terminal to input the required data. The data stream received over the satellite will contain certain bytes marked with a special code to indicate that they may be used for local injection. The local computer recognizes these bytes and, if it has information to send, will replace these bytes by the required information before forwarding the data stream to a signal shaper and SCA generator.

The receiver consists of a specially designed FM receiver with an SCA data demodulator. The output from this demodulator is fed into a microprocessor which is programmed to establish frame synchronizations and select the incoming data relevant to the service required.

Transmission Protocol

The transmission protocol will support the following capabilities:

* A large number of low speed (below 300 bps) data streams for newswire type services.
* A single high speed data channel for block-addressed transmission.
* Error correction and data encryption for selected channels without imposing overhead for services not requiring them.
* A simple receiver design for newswire services.
* Local insertion of data streams at FM station.

Spectral Shaping Techniques

There are two techniques under investigation for the modulating method at 9600 bps. These are proprietary in nature, and it is expected that patents will be applied for and granted.

Receiver Design

The important characteristics of the receiver design are:

* Wide dynamic input range
* Synthesized tuning
* Bandwidth to 100kHz
* High tolerance to co-channel interference
* Insensitivity to noise

These characteristics are achieved in a simple design whose major componenets are shown in figure 4.

The data demodulator is controlled by a crystal with hardware implemented frame synchronization circuitry, and a "flywheel" system which maintains synchronization even in the event of temporary loss or mutilation of received data.

RECEIVER INSTALLATION AND MAINTENANCE

The NIU FM SCA receiver provides the user with a local means of receiving information from the INC network via an FM broadcast station. It is light, compact, can be placed on a desk or shelf, and plugged into a standard 117 VAC power outlet.

The radio signal comes from the antenna into the antenna input of the receiver, using a standard type F connector (the same type used for cable TV connection). The receiver may be permanently tuned to the FM station of a particular city by a quartz crystal. This crystal would have to be changed and the receiver realigned before it can be used in another city. Alternative receiver designs are being considered as part of an ongoing R&D program, in order to eliminate the need for realignment. The receiver decodes and sends subchannel data to a microcomputer controller, which monitors the data for errors. if an error is detected, or in cases of very poor reception a special error signal will be given. The microcomputer also continuously monitors incoming data for messages addressed to it. The address, which is unique to each receiver, is recorded in a removable memory chip which is located in the receiver. Thus the design of the receiver makes it a simple, easily installed and maintained unit.

Installation of receivers will be performed by a field installation and maintenance organization. It is presently anticipated that this service will initially be acquired from a third party by INC Telecommunications on a contract basis, but will eventually become the responsibility of Local Information Utilities. Such a procedure is made possible by the simplicity of the installation, which consists of confirming the availibility of a power supply, installing the antenna (e.g., J-Stub); and performing a check of the signal reception. The receiver is designed to allow for simple, procedural testing.

Maintenance of receivers is likewise a simple process consisting of replacing a board or the entire unit. This replacement would come from local maintenance offices. Faulty units would be repaired at a central location; the appropriate notification of the recceiver swap is made to the central network control by the maintenance organization. Investigations are underway to determine means of automating this process. This process is economical and encumbers the end user with a minimum degree of inconvenience or service delays.

MARKETS

The opportunities existing in the information technology market have attracted nearly every large corporation involved in communications in any way (AT&T, RCA, GTE, etc)., as well as spawning a number of new competitors - SBS, MCI, American Bell, etc. NIU has targeted a very specialized niche in this market with a unique network not well served by the large comporations named above., which will enable it to capture some business from existing competitors., as well as create new demand. It is this specialized niche which forms the basis of NIU's 5 year marketing strategy business plan.

The INC network is designed to provide the lowest cost "last mile" distribution., a service now available from only one source - the telephone companies - and often representing the most expensive leg of a communications network. Furthermore, the cost of using a local telephone loop is expected to rise dramatically, while other services such as cable, multipoint distribution service, and direct broadcast satellite will not be widely available for quite some time. This is happening when demand for (and ability to receive) data is skyrocketing., a tremendous amount of information is being stored in computer data bases., and the pace of change demands that users be informed on nearly a real time basis.

The low cost of point to multipoint distribution using the INC network will allow it to displace networks of leased telephone lines in certain applications such as a newswires. These economics in combination with the features - simultaneous communications, high reliability, easily portable (and hence installable) receivers, and adressability., will also create brand new markets. Customers will use the INC network in ways which the telephone system., telegraph., telex., mail., or messengers could not serve., alone or in combination. Uses of the network are can be grouped into four major categories.

News

News is a nearly continuous transmission of general news items during the day or the smaller daily distribution of specialixed news items, usually at low speed (300 bps). The newswires of UPI, AP, Reuters, and Dow Jones of the many specialized financial or political newsletters are the major examples. These services are distributed on a national basis and generally must reach the subscriber instantly.

Because the major news services now use leased telephone line networks, the primary advantage of replacing them with the INC network is cost saving.

For newsletters now distributed via mail, use of the INC network offers the new cspability of instant and economic communication.

Electronic Messaging

This is a broad class of applications that are characterized by event-driven, sporadic transmission of short messages. Examples include:

* Administrative messages - e.g., policy or organizational announcements.

* Financial results - regular reporting of sales, earnings, inventory, etc.

* Security applications - stolen check or bad credit card notification.

* Emergencies - notification of emergencies for public agencies and others.

These services will be both local and national, and most likely transmitted at "medium" speed - 1200 bps. Some applications will require "priority" service, i.e., a message would not be inserted into the queue at the uplink or transmitter, but would be transmitted immediately.

The value of electronic messaging to the user lies in the ability to take action based on timely information, rather than lower communications costs. Although it is likely that the cost saving realized simply through reduced use of conventional channels could pay for a customer's use of the INC network, the time and effort saved in sending messages to multiple recipients are expected to be much larger benefits.

Downloading

Downloading is a term used to deliver a number of applications; these can be reduced to two basic ones. The first is the modification of an existing data base or program, usually on a regular basis. Examples of this use include regular price and schedule updates such as for grocery chains or airlines, and software maintenance where "patches", (corrections for errors discovered in programs) are disseminated to a software producer field organization or directly to customers.

The second application is the distribution of new information or software. This includes downloading of small databases, such as inventories, customer lists, and routing information, as well as distribution of software, particularly for personal computers where the programs are small.

Downloading applications will generally involve sending data and software directly into computer memory. Because there is no visual inspection of the information prior to its use, particularly high reliability will be required.

The value of downloading lies in distribution efficiency. Use of the INC network in lieu of copying and mailing software or data on discs or tape results in lower costs and faster service. In the case of software maintenance a large benefit could be realized when the timely distribution of a patch avoids a customer complaint and service call.

Remote Operations

The INC network can be used to activate switches and other electronic devices without human intervention. The ability to turn a wide variety of devices on or off while unattended, or start and stop various services, is valuable in many applications.

In particular, many electric utilities are planning major load management programs which include residential appliance control (water heaters and air conditioners). These programs will require hundreds of thousands of control points in homes - a function which the INC network will perform very effectively.

VALUE

Any organization needing to send information to many points on a frequent basis is a potential customer. This includes: news services, newsletters, banks, retailers, franchisers, manufacturers, etc.

Any customer may well use the network in more than one of the ways described - realizing substantial savings over existing systems. However, the greatest value of the network may lie in brand new applications made feasible by the unique capabilities of the INC network.

FIGURE 1

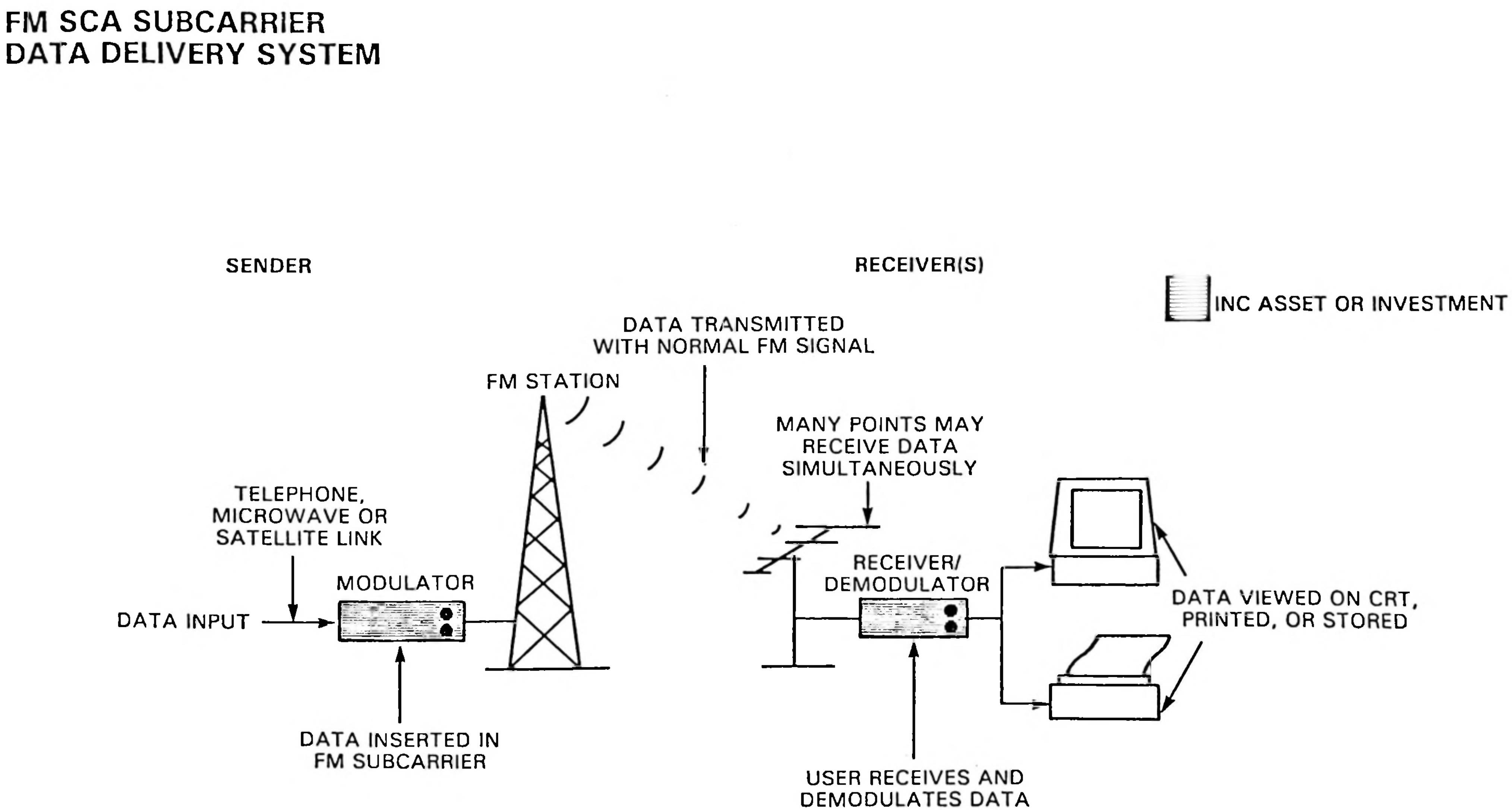
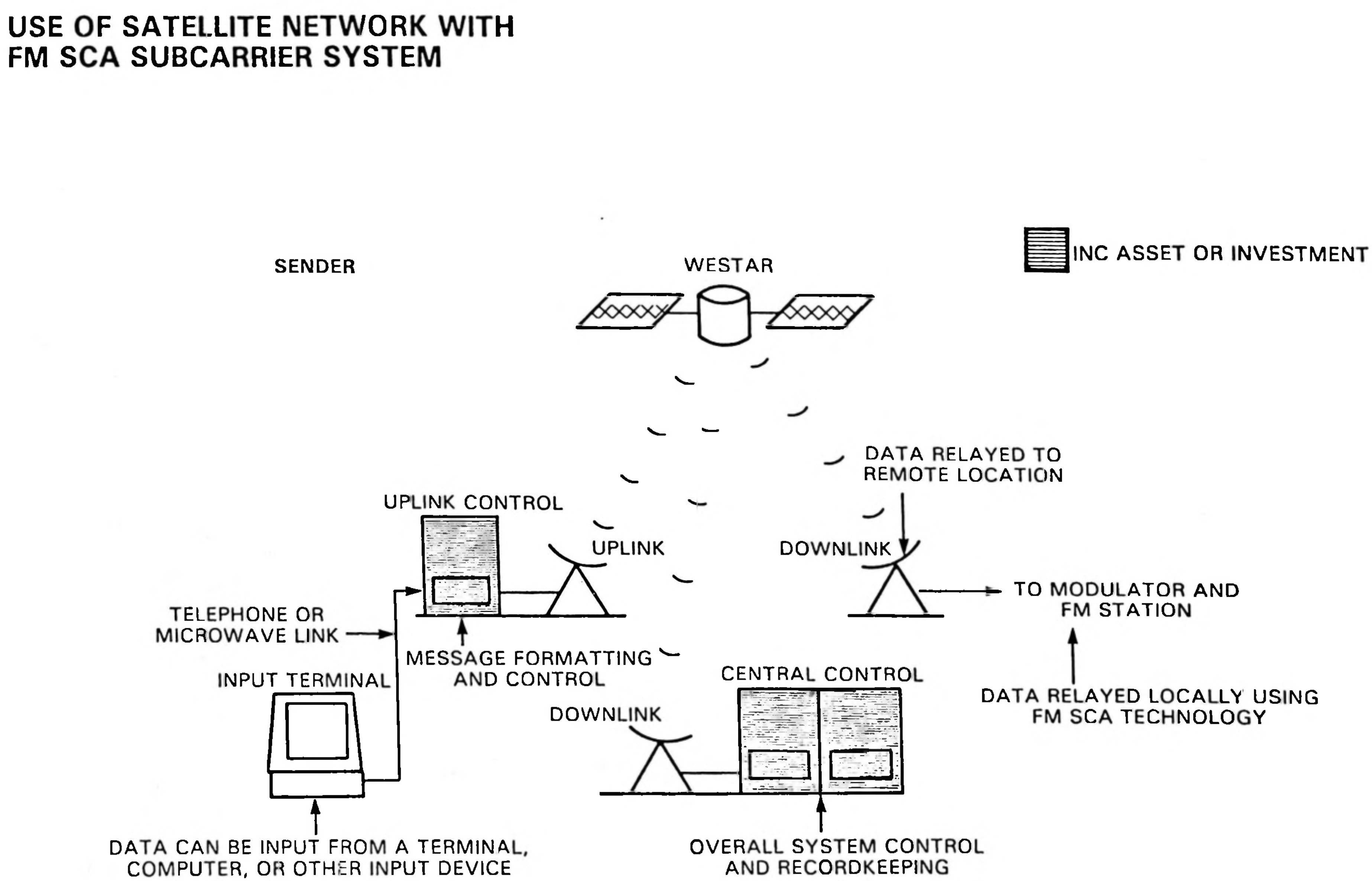

FM SCA SUBCARRIER
DATA DELIVERY SYSTEM
SENDER
RECEIVER(S)
INC ASSET OR INVESTMENT
DATA TRANSMITTED
WITH NORMAL FM SIGNAL
FM STATION
MANY POINTS MAY
RECEIVE DATA
SIMULTANEOUSLY
TELEPHONE,
MICROWAVE OR
SATELLITE LINK
MODULATOR
RECEIVER/
DEMODULATOR
DATA INPUT
DATA VIEWED ON CRT,
PRINTED, OR STORED
DATA INSERTED IN
FM SUBCARRIER
USER RECEIVES AND
DEMODULATES DATA

FIGURE 2

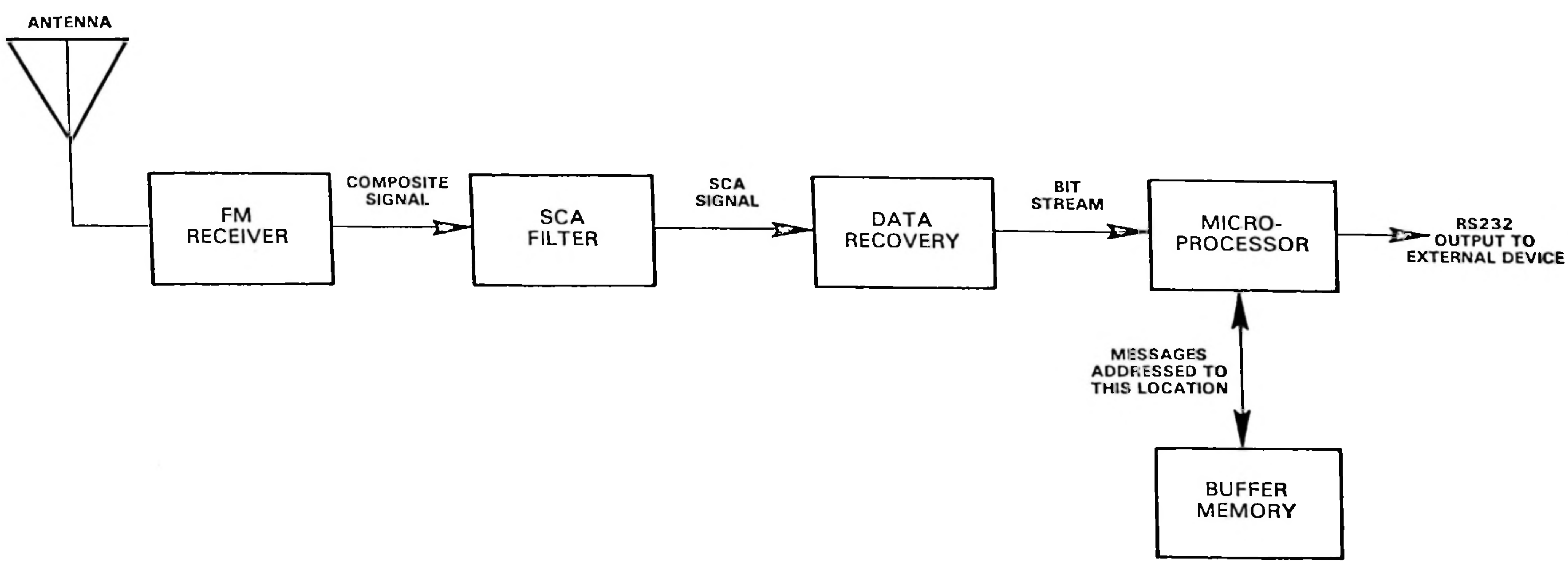


USE OF SATELLITE NETWORK WITH
FM SCA SUBCARRIER SYSTEM
SENDER
WESTAR
INC ASSET OR INVESTMENT
DATA RELAYED TO
REMOTE LOCATION
UPLINK CONTROL
UPLINK
DOWNLINK
TO MODULATOR AND
FM STATION
TELEPHONE OR
MICROWAVE LINK
MESSAGE FORMATTING
AND CONTROL
INPUT TERMINAL
CENTRAL CONTROL
DOWNLINK
DATA RELAYED LOCALLY USING
FM SCA TECHNOLOGY
DATA CAN BE INPUT FROM A TERMINAL,
COMPUTER, OR OTHER INPUT DEVICE
OVERALL SYSTEM CONTROL
AND RECORDKEEPING

FIGURE 3
NATIONAL PUBLIC RADIO STATIONS

FIGURE 4

FUNCTIONAL BLOCK DIAGRAM
OF DIGITAL FM SCA RECEIVER



IMPROVING THE SIGNAL-TO-NOISE RATIO

AND COVERAGE OF FM STEREOPHONIC BROADCASTS

EMIL TORICK

CBS Technology Center

Stamford, CT

ABSTRACT

It is a well-known phenomenon that the presently authorized system for FM stereophonic radio broadcasting imposes a significant degradation of the received signal-to-noise ratio when such broadcasts are compared with simple monophonic transmission. As a consequence, station coverage area is reduced both for monophonic and biphonic reception. However, recent studies have demonstrated that this degradation can be eliminated by the use of audio companders, and this paper analyzes the potential realizable improvements. It also shows how such companding may be implemented in the broadcast signal without compromising existing service.

INTRODUCTION

The potential of FM sound broadcasting as a high fidelity medium has been recognized since the earliest commercial FM radio broadcasts in the 1940's. Because of its relative immunity to electromagnetic interference and the ability to provide full audio bandwidth with low noise, frequency modulation was also selected as the transmission method for television sound. Although FM radio was hardly a universal success in the commercial sense when stereophonic broadcasts were first authorized in 1961, it was not long before the attraction of two-channel high-fidelity sound helped elevate FM to the status it enjoys today. But, although stereophony adds a new acoustical dimension to radio reception, it does so only at the expense of a serious degradation of another high fidelity parameter — the signal-to-noise ratio.

This noise penalty in stereophonic broadcasting is well known. However, less obvious is the restrictive influence this phenomenon has on station coverage. Often, stereophonic reception limits experience reduction typically to one-fourth or one-fifth the area of simple monophonic broadcasts.

THE NOISE PENALTY

Several factors contribute to the higher noise levels and coverage losses resulting from multichannel transmissions. When a station converts to biphonic* service, monophonic coverage is reduced since signal power must be divided among the various components of the more complex baseband signal. The biphonic signal-to-noise ratio is less than monophonic signal-to-noise ratio because of the wide bandwidth of the composite signal. The equation of this signal is familiar:

$$f(t) = M + p \sin \omega/2 \; t + S \sin \omega t$$

where M is the monophonic sum signal, p is the pilot, S is the stereophonic difference signal, and $\omega = 2\pi \, 38\text{kHz}$. With a baseband spectrum extending to 53 kHz for biphonic transmissions, the noise level is particularly high because of the rising spectral characteristic due to frequency modulation. As illustrated in Figure 1, the so-called "triangular" noise spectrum increases 6 dB/octave with increasing frequency of the composite signal. Audio de-emphasis counteracts this somewhat, as shown in this figure, but the noise problem is still severe. After demodulation, the noise components of the difference channel subcarrier are added, statistically independent, to the noise already present in the monophonic signal during audio dematrixing.

A precise computation of the theoretical loss of signal-to-noise ratio must take into account factors such as the effect of de-emphasis, the format of the audio test signal which is assumed, and interleaving. Interleaving is an interesting phenomenon whereby with certain audio signals the peak amplitude of the sum of the main and subchannels may be less than the sum of the peak amplitudes of these channels, thus permitting the interleaved signals to to be raised to full modulation, with a resultant improvement in the signal-to-noise ratio. A number of researchers have previously studied these factors.

---

*The term "biphonic" is used throughout this document in order to clearly differentiate two-channel broadcasting from other forms of stereophony such as triphonic and quadraphonic broadcasting.

A calculation of the signal-to-noise degradation in biphonic broadcasting was published by Parker and Ruby[1] in 1962. In it, Parker and Ruby assumed the transmission of the peak monophonic power available, i.e. no modulation of the subcarrier. (L - R = 0) While their conclusion of 23 dB degradation has received widespread acceptance, this figure is not completely representative of typical programming. More recently, under EIA auspices, the National Quadraphonic Radio Committee (NQRC) studied the subject in greater detail. In its final report[2], the NQRC reaffirmed the 23 dB penalty for a monophonic test signal, but also, by using a wide variety of audio test signals, demonstrated that 26 dB is more representative of stereophonic programming with wide audio separation. For monophonic receivers, the NQRC data predict noise degradation of 1 to 7 dB, depending on the particular type of test signal used.

Such losses of signal-to-noise ratio also cause a reduction in the effective area of coverage for a broadcast station. Figure 2, based on NQRC data[3], illustrates this effect for a representative set of transmission and reception conditions*. For reception at a 50 dB signal-to-noise ratio, the limit of station coverage would extend to a radius of 128 miles when monophonic transmission only is employed. However, with biphonic transmission, monophonic reception is reduced to 100 miles, and two-channel reception extends only to a 60-mile radius. Although in reality station service areas are often limited by co-channel and adjacent-channel interference rather than by noise, Figure 2 represents a useful comparison of the theoretical limits.

NOISE REDUCTION BY COMPANDING

The application of companding systems offers a potential solution to the problem of the noise penalty. Companding systems achieve noise reduction by compressing the dynamic range of an audio program before transmission and expanding it to its original dynamic range at the receiver. Figure 3 illustrates this effect. On the left, the "original program" signal, with a wide dynamic range and a low noise level is represented. In the center, for this example, the program is compressed to approximately one half its original dynamic range for transmission. During transmission, additional noise is introduced at a level below that of the compressed program, but at a level which would have intruded on the program had it not been compressed. Finally, the "expanded program" is shown reconstituted to its original dynamic range and with the transmission noise simultaneously reduced to an inobtrusive level.

Companding systems have achieved success in various audio applications, including tape and disc recording. For broadcasting, tests were conducted in Sweden in the early 1960's utilizing a companding system in the S channels of FM-AM and FM-FM transmission systems. Favorable results were reported for the FM-FM transmissions, although the system was never fully implemented. During the last 20 years, significant improvements in companding systems have been achieved. Renewed interest in broadcast applications is now evident in the industrywide studies of

---

*The NQRC used the FCC FM Engineering charts for the estimated field strength exceeded at 50% of the potential receiver location for at least 50% of the time, with a dipole receiving antenna height of 30 feet. The transmitter height was assumed to be 1000 feet, with a 10 kilowatt effective radiated power at 98 MHz. The receiver assumed to have a 10 dB noise figure.

methods for transmitting multichannel sound in television broadcasts[4]. Under EIA and NAB auspices, the Multichannel Sound Committee of the Broadcast Television Systems Committee is examining the potential application of companders to the S channel for television audio. Although the studies are not yet complete, preliminary results seem to indicate potentially acceptable performance by at least four proponent systems. (CX, DBX, Dolby, and Hy-Com)

COMPANDERS FOR FM RADIO BROADCASTING

Given the recent advances in the art of audio companding, it is once again appropriate to examine the potential application to FM radio broadcasting. At the present time, some broadcasters utilize Dolby-type encoding to provide modest noise reduction in receivers equipped with appropriate expanders and, through simple receivers without expanding capability, relatively acceptable playback. However, the need to maintain this compatibility with simple receivers inhibits the potential for truly significant noise reduction in the other (more capable) receivers.

There is an alternative approach that can achieve both better noise reduction and full compatibility with existing receivers and which avoids any modification of the audio signals in either the M or S channels. In this scheme a companded S' channel would be transmitted, additionally, in order to provide noise-free biphonic reception to a new class of receivers. (Existing receivers would continue to utilize the conventional S channel.) The new S' channel can be transmitted in quadrature with the existing stereophonic subcarrier as follows:

$$f(t) = M + p \sin \omega/2\ t + S \sin \omega t + S' \cos \omega t$$

Figure 4 represents the baseband spectrum of the suggested new service. The quadrature subcarrier requires no additional spectrum space, and, as will be shown later in this paper, imposes only a negligible penalty in modulation potential.

Figure 5 is a block diagram of an encoder for the new composite signal. A conventional audio matrix is used to derive the sum signal M and the difference signal S. A carrier generator provides the 19 kHz pilot tone and the sine and cosine functions of the 38 kHz subcarrier. The conventional S channel modulates the sine carrier and the compressed difference signal S' modulates the cosine function. The M, S, S' and pilot signals are added together to constitute the composite signal. In the receiver (Figure 6), a switch is employed to select either the S or the S' channel (expanded) for dematrix purposes. Signals for automatically actuating the switched functions either by amplitude modulation of the pilot tone or the addition of a separate identification signal have previously been suggested[5], and could be considered for this service.

PERFORMANCE IMPROVEMENTS

The effect on signal-to-noise caused by the addition of a new compressed-biphonic channel is compared with the performance of conventional monophonic and biphonic transmissions in Table 1. Based on NQRC calculations, the table shows the predicted performance for various combinations of the three transmission and receiving modes. Two test signals are shown — a L + R signal equivalent to that used by Parker and Ruby, and L (or R) only, representative of most of the NQRC calculations. For either modulating signal, compressed-biphonic reception can be as good as equivalent monophonic reception if an ideal companding system is employed, i.e., sufficient noise reduction is achieved in the S' channel to allow the noise of the M channel to predominate. In the previously referenced study of multichannel sound for television[4], specifications of proponent systems call for a combination of audio companding and an S channel modulation level which is twice that of the M channel. Such a higher subcarrier level is not recommended for radio service because of the additional penalty it would impose on monophonic and conventional biphonic reception.

A prediction of the reception range limits for a 50 dB signal-to-noise ratio with companded biphonic transmission is shown in Figure 7. (The NQRC method and the FCC 50, 50 charts[6] were used.) In comparison with conventional biphonic transmissions, the new companded system causes a relatively insignificant reduction of monophonic reception from a 100 mile to a 96 mile radius, and a similarly modest reduction of biphonic reception from 60 miles to 56 miles. However, the new companded biphonic service would extend all the way to the monophonic contour at 96 miles. This represents approximately a three-fold increase in coverage area over the existing biphonic service, as compared with existing service.

An alternative way of displaying the signal-to-noise of various transmission schemes was also suggested by the NQRC and is illustrated in Figure 8. Signal-to-noise ratios are shown at the so called "urban" contour (E_0 = 1mv/M) and the "rural" contour (E_0 = 50 v/M). For the conditions assumed (which are the same as for the previous figures) all systems will exhibit the same signal-to-noise performance at the urban contour. With such high field strength, reception characteristics here will be dictated only by receiver performance; typically, a signal-to-noise ratio of 65 to 70 dB may be realized. At the rural contour, (70 mile radius), conventional biphonic receivers will exhibit a 43 dB signal-to-noise ratio, while companded biphonic receivers will achieve 62 dB.

CONCLUSIONS

Sufficient evidence exists to warrant further consideration of the adoption of new companded service for FM radio broadcasting. As shown here, such transmission has the potential for providing service nearly equivalent to that provided by monophonic receivers.

Compatibility with existing receivers can be maintained if the compressed signal is encoded in a new channel which is in quadrature with the S channel.

The implementation of appropriate transmission equipment is easy to achieve.

For receivers, commercial integrated circuits for expanders are already available in the marketplace, and the technology for decoding the S' channel has previously been developed for other applications such as AM stereo and FM quadraphonic reception.

Finally, a sufficient number of presently dissatisfied radio listeners will provide a ready market for the new class of receivers.

ACKNOWLEDGMENT

The improved FM broadcast system described here was jointly conceived by the author and Mr. Thomas Keller of the National Association of Broadcasters. The author is grateful for Mr. Keller's encouragement in the preparation of this paper.

REFERENCES

1. N. Parker and D. W. Ruby, "Some notes on the calculation of the S/N ratio for a FM system employing a double sideband AM multiplexer signal," IEEE Trans. Broadcast Telev. Receivers (International Convention Issue), vol. BTR-8, pp. 42-46, Apr. 1962.

2. J. Gibson, et al, "Signal and Noise Levels in Proposed Multiplex Systems For FM Broadcasting of Quadraphonic Sound," Report of the National Quadraphonic Radio Committee to the Federal Communications Commission, Vol. II, Chapter 1, Section 1.4, Nov. 1975.

3. J. Gibson, et al, "Illustrations to Relations Between Signal-To-Noise Ratio and Range in Existing FM Broadcast Services and Proposed Systems For FM Broadcasting of Quadraphonic Sound," Supplement to Volume II, Chapter 1, Section 1.4 of the NQRC Report, Jan. 15, 1976.

4. EIA-BTS Committee, "Multichannel Television Sound: The Basis for Selection of a Single Standard", Vol. 1, National Association of Broadcasters, July 16, 1982.

5. E. Torick, "A Triphonic Sound System for Television Broadcasting", 124th Technical Conference of the SMPTE, Nov. 1982.

6. FCC Rules and Regulations, Volume III, Part 73.333.

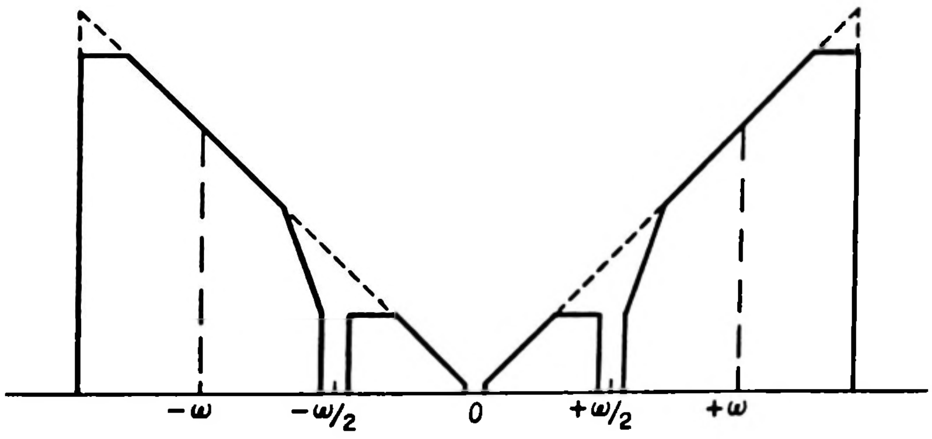


FIG. 1- FM NOISE SPECTRUM, WITH DE-EMPHASIS

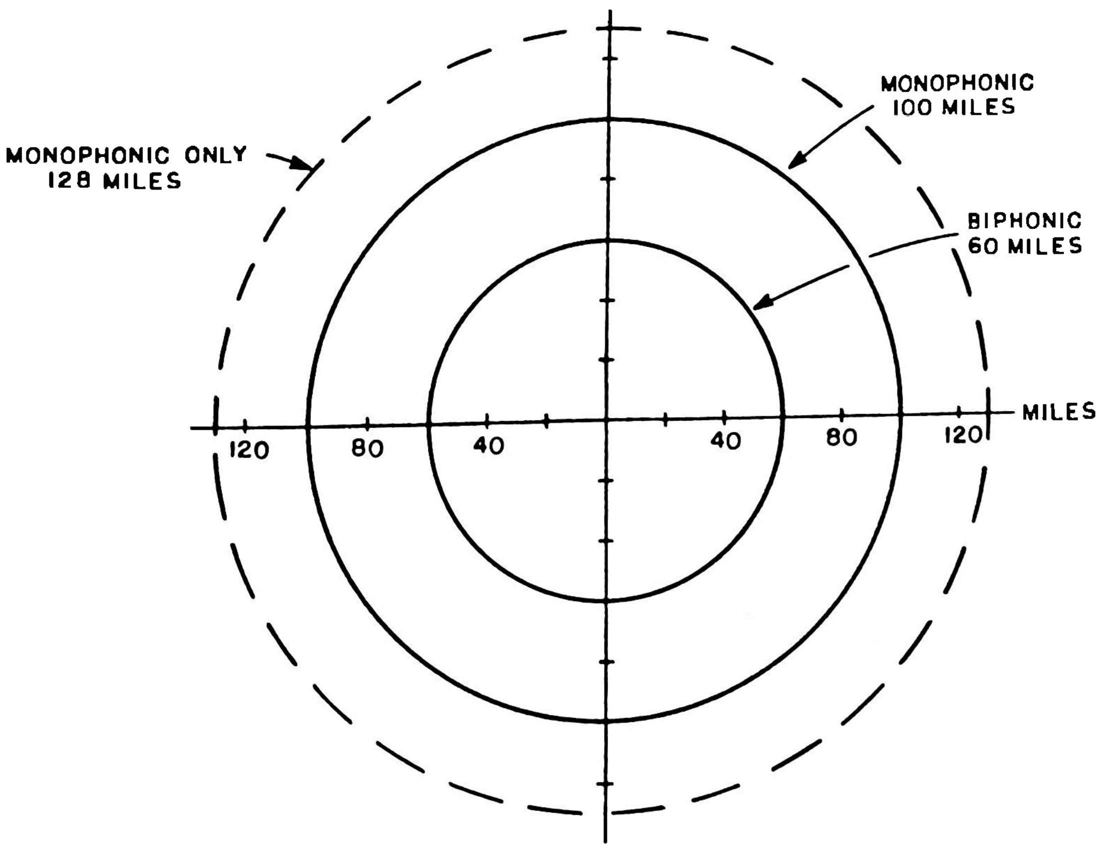


FIG. 2 - RECEPTION LIMITS FOR 50 dB SNR

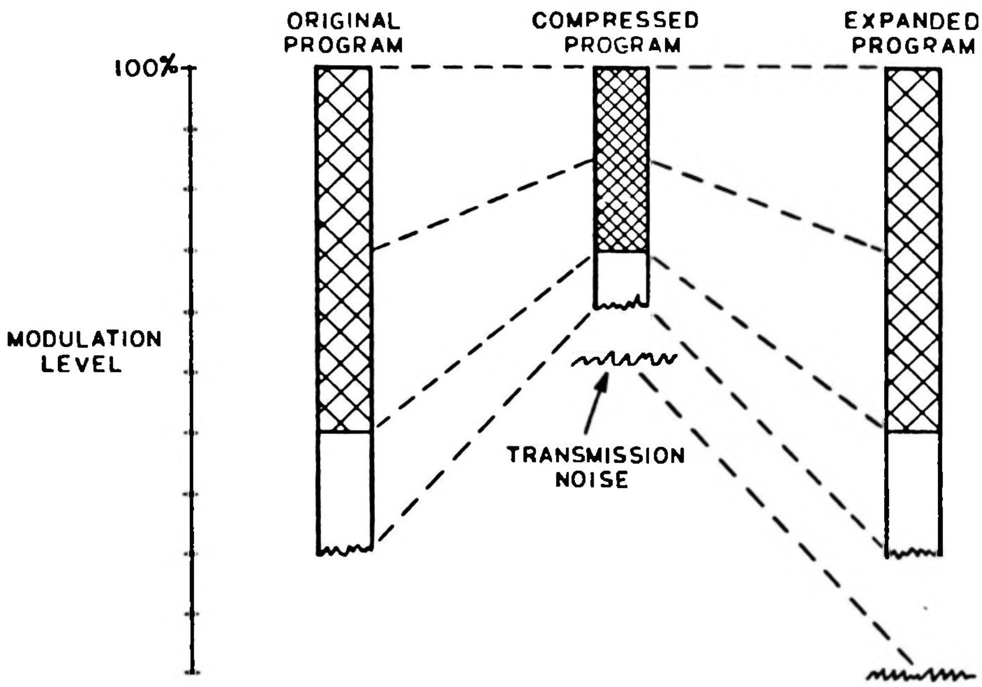


FIG. 3- COMPANDING

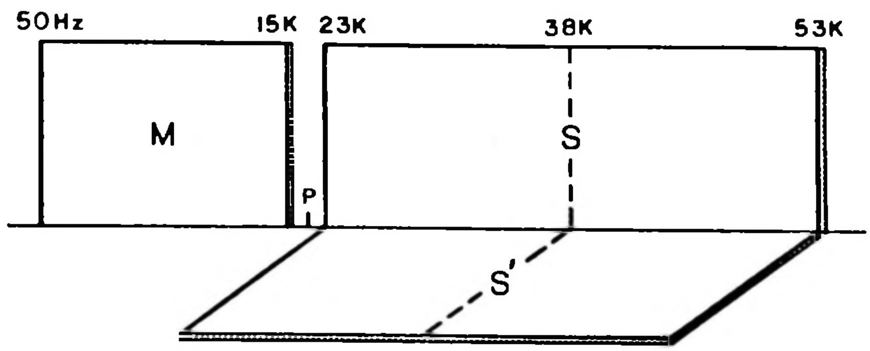


FIG. 4- BASEBAND SPECTRUM

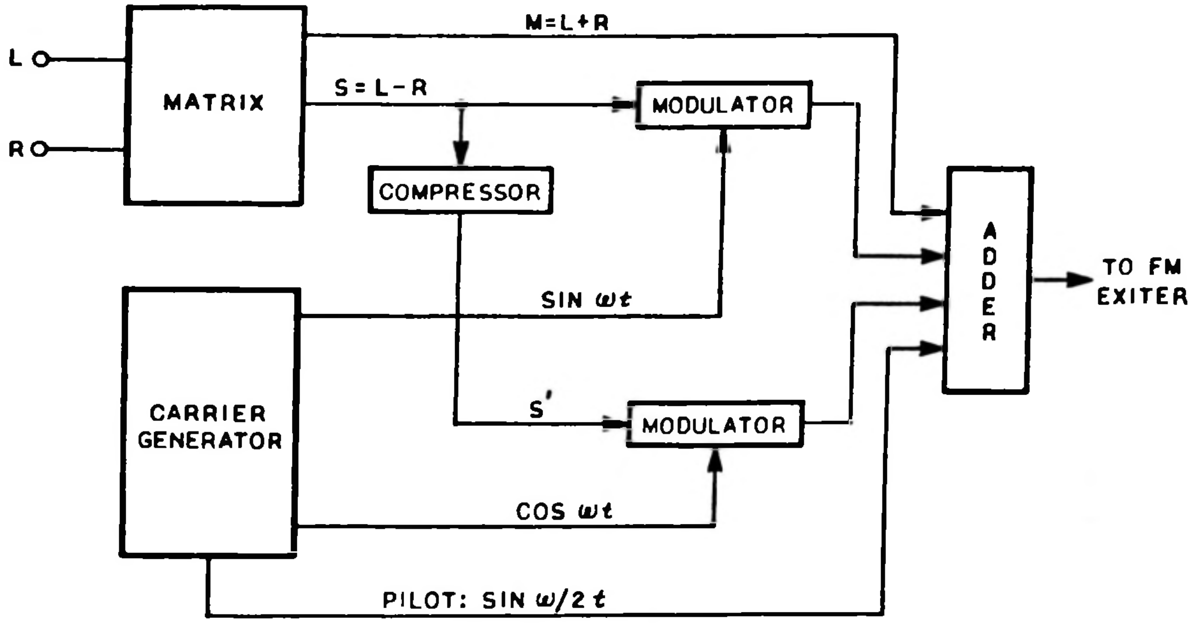


FIG. 5- ENCODER

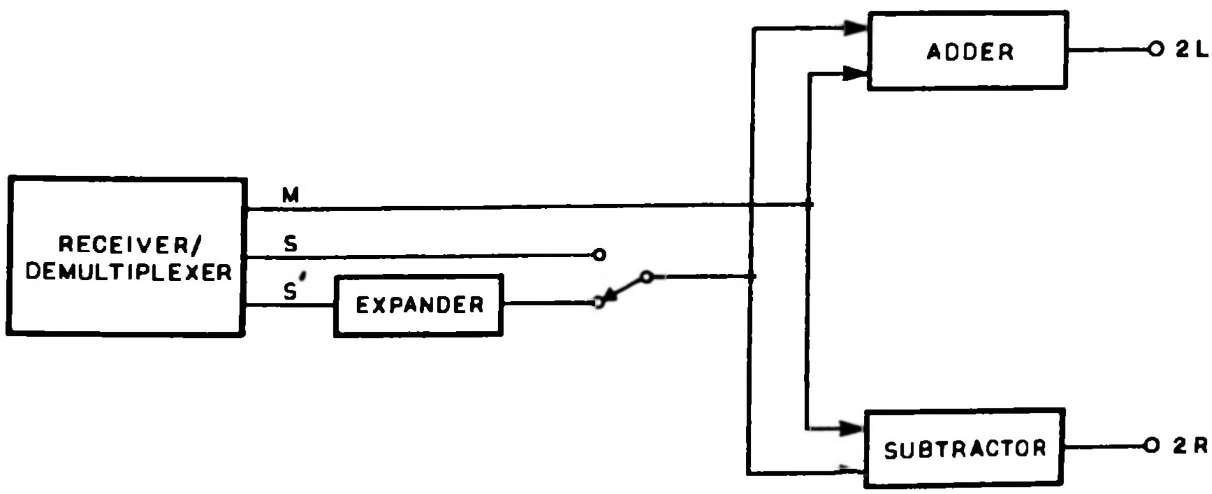


FIG. 6- DECODER

Modulating Signal	Transmission Mode	Receiver State		
		Monophonic	Biphonic	C-Biphonic
L+R Only	Monophonic	0	N. A.	N. A.
	Biphonic	0.92	23.11	N. A.
	C-Biphonic	0.92	23.11	0.92
L (or R) Only	Monophonic	0	N. A.	N. A.
	Biphonic	6.94	26.12	N. A.
	C-Biphonic	8.58	27.76	8.58

Table 1. Relative Receiver Signal-to-Noise Degradations (dB)

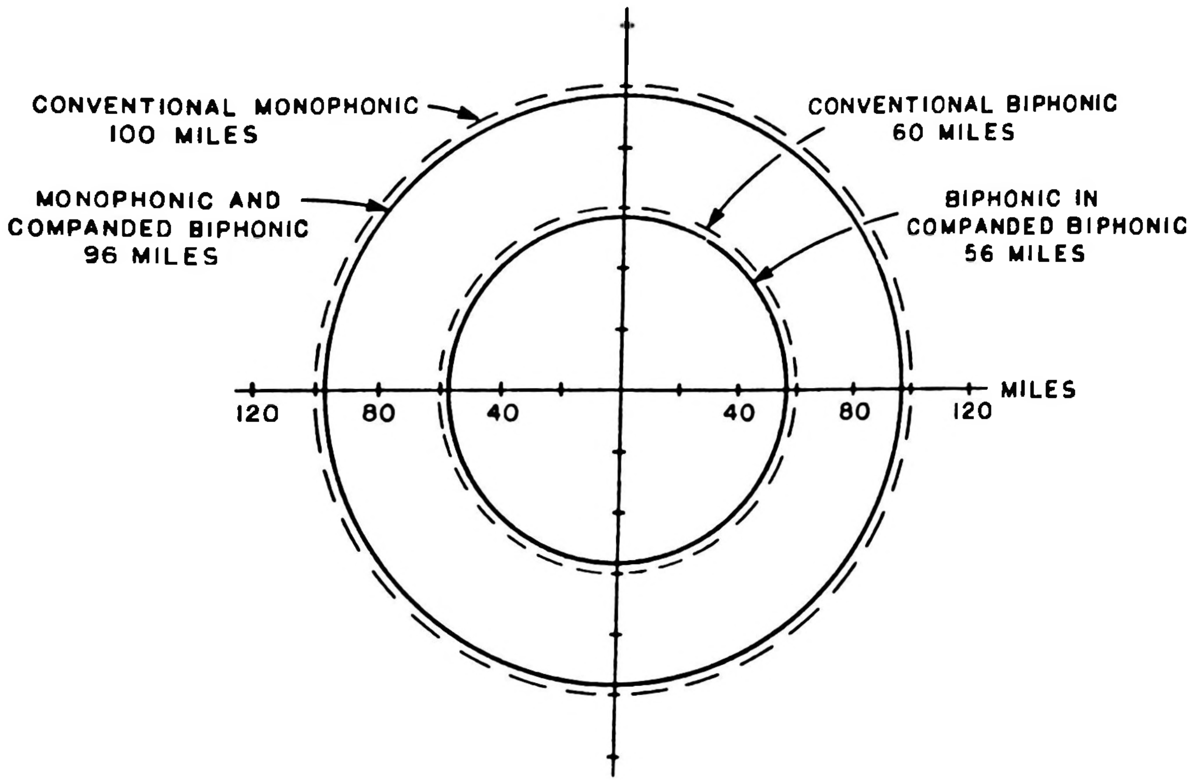


FIG. 7- RECEPTION LIMITS FOR 50 dB SNR

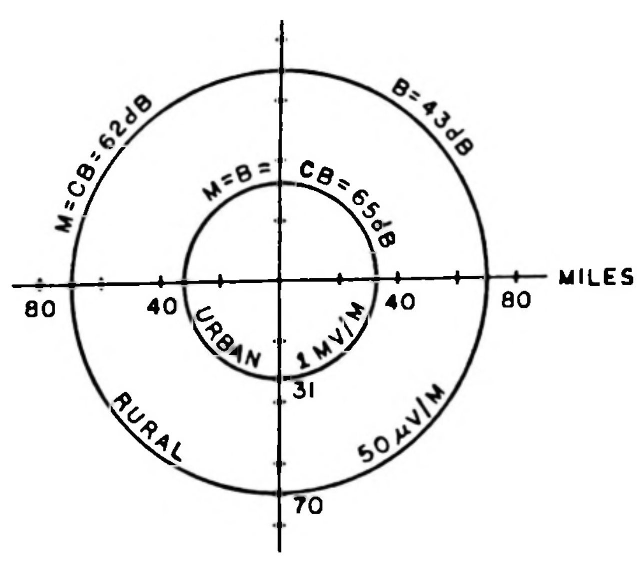


FIG. 8 - SIGNAL-TO-NOISE RATIOS AT KEY CONTOURS

Satellite Digital Audio Transmission: An Overview

Edmund A. Williams

National Association of Broadcasters

Washington, D. C.

* This paper was originally presented in part at the September 1982 IEEE Broadcast Symposium in Washington, D. C. while the Author was with the Public Broadcasting Service.

Satellite Digital Audio Transmission:
An Overview

Since 1978 when 15 kHz audio began accompanying television programs as a regular service replacing the old 5 kHz audio, the broadcasting industry has discovered that the public not only demands better quality audio, but that the latest transmission technology, digital, was becoming available to provide this service.

High quality audio transmission is by no means new, it just has not enjoyed the ease of transmission made available by satellites and by digital techniques. In the past, although it was indeed possible to order a 15 kHz audio circuit from the terrestrial common carriers, the problems with the circuit and the process, not to mention the cost (which was based on a base rate plus a mileage charge) sometimes made acquiring "hi fi" sound not worth the effort. If the program was in stereo, two of these circuits were needed and you hoped they both took essentially the same path so as not to disturb the stereo effect too much.

The effect the difficulty of obtaining and maintaining wire based high quality audio circuits created, was to fall back on the 5 kHz circuits which, although not the best sounding, were reliable because they had been in place since the hey-days of network radio and were reasonably well understood by the carriers. Network television audio inherited many of these circuits which were used until 1978 when AT&T added a subcarrier to their video channel to carry high quality audio. While this benefitted television stations, the radio networks still were saddled with the 5 kHz circuit.

It is the era of the satellite however, that makes it possible to distribute high quality audio easily as well as at a reasonable cost. Distance is no longer a factor and the fact that there is only one "repeater" in the satellite system substantially reduces the chances for signal deterioration. A satellite transmission system is in essence a two-hop microwave circuit and, for all practical purposes, can be treated as just another element in a multihop microwave transmission system. For a common carrier serving many diverse customers, the satellite circuit might reduce the number of repeaters from several hundred to less than a dozen. For the broadcaster, there may be as few as one repeater (the satellite) or as many as 4 or 5 when including the microwave interconnecting circuits between the satellite earth terminals and the source (network) and user(station).

With the satellite able to handle the kind of microwave circuits familiar to broadcasters, the next step was to add analog audio subcarriers to video circuits for 15 kHz audio service. Typically, two are added for stereo and several others can be added if some degradation can be accommodated in either the sound or picture signals. But there has been a need for more subcarriers to accomplish a number of diverse audio requirements. Multiple audio channels for television will be needed for stereophonic sound, special monophonic mixes, narration, foreign language and non-program related sound circuits for cable TV systems and FM radio stations. For most of these applications analog transmission technology utilizing FM subcarriers has been considered adequate and, given that the audio signal is not remodulated too often, will meet the audio needs of most broadcasters.

But the future belongs to digital technology and, when planning for satellite audio transmission in the 80's and 90's, it is appropriate that digital transmission be given serious consideration. Digital techniques are now understood to the extent that world-wide digital standards are now beginning to emerge. Audio recording equipment manufacturers have only recently agreed that a 48 kHz sample rate and 16 bit quantization levels are acceptable values which will eventually permit international exchange and allow the orderly development of editing and audio mixing and processing equipment.[1]

Digital audio transmission however, does not enjoy such a degree of standardization probably because of the nature of the interests of the users of the digital audio circuits. Although the sample rates and quantization levels have been more or less agreed to at 32 kHz and 12 bits respectively, major differences exist in the transmitted bit stream which will not permit receivers to be interchanged between the various systems (something which can be done with most of the analog systems). A combination of evolving technology and specific user requirements has produced only a semblance of standardization in digital audio transmission although there are industry groups actively involved in establishing digital transmission standards. Perhaps a review of several existing and planned digital audio systems will explain how these systems have evolved and why it is unlikely that digital audio transmission standards will ever be completely compatible.

In February 1978, the Public Broadcasting Sevice (PBS) inaugurated its first phase of their nationwide television distribution system via satellite.[2 3 4] In addition to the usual FM audio subcarrier at 6.8 MHz in the video baseband, there was a four-channel digital audio subcarrier placed at 5.5 MHz. (Figure 1). The system, developed jointly by PBS and Digital

Communications Corporation between 1974 and 1978, was called DATE (Digital Audio for TElevision).

The DATE system was originally developed to be used on long haul terrestial microwave circuits commonly used by television networks.[5,7] The DATE system was specifically designed to work with the AT&T TD series television terrestrial microwave transmission facilities and was able to accommodate the kind of distortions found in multi-hop microwave systems. The Public Broadcasting Service, which used AT&T facilities at the time and which distributes a substantial amount of music-oriented programming to over 280 public television stations throughout the US. PBS was vitally concerned that the quality of audio (5 kHz) provided with the television circuit was not adequate for the high fidelity transmission requirements of music programs. While television studios, recording equipment and broadcast facilities were all able to handle 15 kHz audio, the interconnection link between program source and broadcaster could not. At the time the digital audio system was being considered AT&T had not yet fully developed its FM audio subcarrier system and an earlier-than-planned implementation of the FM system would have been too costly.

The DATE digital parameters were selected based on certain known and assumed factors. For example, the sample rate must be greater than twice the maximum frequency to be transmitted (15 kHz) and should be related to TV color subcarrier (3.58 MHz) to permit a possible interface with video components. The sample rate selected was 34.42 kHz (3.58 MHz ÷ 104). Four channels were needed to provide stereo plus mono plus a spare channel for future uses. The placement of the DATE QPSK 1.79 Mbit carrier at 5.5 MHz was determined by a careful analysis of the spectrum of the AT&T TD-2 terrestial microwave system. (Figure 2). The subcarrier is strategically located just above video and just below the second harmonic of the color subcarrier. The system was never implemented on terrestial facilities due to the development of the satellite system on which it is now in daily use.

In order to achieve a satisfactory bit error rate (BER) on the DATE system certain extra features were added to the normal digitizing process. The 14 bit quantization levels were companded to 11 bits to which a parity bit and compander law bit were added for a total of 13 bits transmitted. The result is 13 bits x 34.42 kHz sample rate x 4 channels = 1.79 Mbit. Using QPSK (quadrature or 4 phase shift keyed), the occupied bandwidth is also 1.79 MHz (QPSK = 1 bit bit/Hz). The carrier is placed at 5.5 MHz on the video baseband at a level of about 16 dB below peak video in order to prevent intermodulation products from being visible in the video signal. In an effort to keep the BER at a value which would be acceptable (10^{-7})[6], some modest error concealment circuitry was included. For example, a detected parity error caused the system to reuse other valid samples and a continuous string of errors produced a gentle muting of the audio until the error rate returned to normal levels. The overall effect is to achieve apparent 10^{-7} error performance with an actual 10^{-5} error circuit or a decoding improvement about 3 dB. This improvement reduces the required video signal carrier-to-noise ratio from about 14 dB to 11 dB. The 14 bit quantizing level was a compromise between significantly higher cost 16 bit analog-to-digital (A/D) converters and lower quantizing levels at which point distortion became discernible.[7] More extensive error concealment or correction circuitry and higher level quantization could have been incorporated, but a much higher cost. Today both 16 bit A/D converters and error correction circuits, able to withstand bearer channel BER of 10^{-4} or less, are available at moderate cost and are now standard design items in digital audio transmission systems.[8] Nonetheless, the DATE system routinely provides excellent frequency response (+ 0.5 dB, 20 Hz-15 kHz), a signal-to-noise ratio of better than 70 dB and total harmonic distortion (THD) levels of less than 0.3%. (Table 1).[9]

Since the development of DATE, other work has taken place in digital audio transmission. Two of the newer systems will be discussed further in this paper because they represent true digital transmission rather than being carried on another signal as is the situation with the DATE system.

The national radio networks (in particular ABC, CBS, NBC and RKO) for the past several years have been considering the use of the satellite for distribution of their programming to affiliated radio stations. There has long been a need for higher quality audio than that provided by terrestrial common carriers particularly if the program material contained music. In the past, AM stations required no more than 5 kHz service provided by the networks. When FM stations became more popular and began relying on the network for more than just news and talk programming, it became apparent that not only was higher quality required, but stereo was needed as well. With the advent of AM stereo, a national service will be required for that as well as for FM.

Radio broadcasters and listeners alike are becoming more sophisticated in their programming tastes and quality of the received signal. Somewhat to the dismay of the broadcasters (and listeners) many listeners now have the capacity to receive a better signal than is frequently broadcast by FM stations. There is considerable catch-up to be done in this area by broadcasters, many of whom are now incorporating recording quality components into studio facilities and improving signal processing and modulation equipment.

Network radio, in responding to the demand by stations and listeners for higher quality audio and to overcome the problems which plague terrestrial distribution sytems, will soon begin to transmit programs to their affiliates in a digital format, via satellite, directly to stations, in effect bypassing the terrestial distribution system. ABC and NBC expect to begin operation via satellite late in 1983, followed by CBS and RKO in 1984. The system they have selected is being built by Scientific-Atlanta (SA) and operated by RCA Americom over the SatCom series satellite. RCA calls their system Audio Digital Distribution System (ADDS). SA will also install several thousand satellite receiving systems at individual stations.

RCA proposed the ADDS to the networks on the basis that it could provide top quality sound, multichannel capacity, a reasonably high degree of security and the use of 3 meter diameter antennas for satellite signal reception at most locations in the continental U.S. The state of the digital audio art will permit these goals to be met with a system providing the following features.

RCA and SA spent considerable effort studying the networks' requirements and how digital audio technology could best be utilized to achieve the goals. A 32 kilobit sample rate is used with 15 bit quantization levels per sample, which are companded to 11 bits. A parity bit is then added to produce (12 bits x 32 K) a 384 kilobit digital data stream. Up to 20 of the 384 Kbit channels can be combined by a multiplexer to produce a 7.68 MHz data stream. Because this stream is vulnerable to transmission problems, which can result in unacceptably high BER, the ADDS system employs a 7/8 rate forward error correction circuit (FEC) (and hard decision decoding) and modulates the carrier as a biphase frequency shift keyed (BPSK) digital signal. The result of using the 7/8 rate FEC is to produce a data stream of 8.777 Mbits. The use of BPSK at this rate causes the occupied bandwidth to be 17.554 MHz or two times 8.777 Mbit (BPSK = 1 bit/2 Hz). This rate was selected because of the lower satellite power density needed to produce the desired BER with 3 meter earth terminals and the fact that BPSK signals are more tolerant of circuit distortions than QPSK signals. The result is a signal which can fit neatly in the satellite transponder frequency spectrum between terrestrial interference sources at ± 10 MHz from the ADDS carrier (Figure 3). Careful bandpass filtering in the receiver will substantially reduce the potential for interference from terrestrial sources, and will allow small earth terminals to be located at virtually any individual radio station.

The combination of parity error detection, 7/8 rate forward error correction and biphase encoding permits the ADDS system to produce a bit error rate in excess of 10^{-7} in a channel that otherwise would produce errors of 10^{-3} or a decoding improvement of about 6 dB. Use of 7/8 rate FEC, BPSK and 20 channel time-domain multiplexing (TDM) also results in a relatively high level of security. A "pirate" would not only have to expend considerable effort to decode the signal, but would need to know the framing code (which establishes the multiplexing order of the 20 audio channels) in order to obtain a useable audio output.

The ADDS, while probably representing an excellent set of compromises of circuit elements and performance objectives, is however, not without certain drawbacks. One of the 20 channels is not completely available for 15 kHz audio, but is normally assigned to certain digital "housekeeping" functions, such as framing codes and other control signals. Also the ADDS signal requires the use of a full transponder due to the power required (32 dBw e.i.r.p.) to provide adequate received signal from 3-meter antennas, even though additional spectrum space in the transponder is available. RCA plans eventually to dedicate two transponders to ADDS for a total of 38 high quality audio channels of 15 kHz each, or other mixes of 15 kHz, 7.5 kHz audio and 4 Kbit data. ABC may use up to 18-15 kHz channels for audio and two other 7.5 kHz channels and cueing and data. NBC is planning to use 6-15 kHz and 4-7.5 kHz audio channels plus others for cueing and data.

Dedicating a full transponder to one ADDS does not permit access to the system from more than one location. Although ADDS is a TDM system, it is not multiple access (TDMA). That is, all 20 channels must originate from the same uplink unless complicated and very costly timing systems are employed to permit access from other uplinks. The networks are not concerned by this because most programming originates at the network head-quarters and infrequent remote progam originations can be accommodated through other facilities (one of which will be discussed later in this paper).

The success of transmitting audio via satellite is based on the reduction of the number of repeaters and the number of times the signal must go through the modulation/demodulation process. The same is true for digital and, if only one encoding section is employed, the least amount of degradation will be produced. To accomplish this, the radio networks will digitally encode the audio signal programming at their respective headquarters in New York and transmit the digital signals to the satellite uplink in New Jersey via microwave, where they will be reconfigured into the ADDS format for transmission to the satellite. For example, NBC will encode six 15 kHz audio channels into six 384 Kbit streams and the for 7.5 kHz channels into two 384 Kbit streams, which along with additional capacity for data, can be combined into two T1 digital circuits1 suitable for transmission over microwave links from the NBC studios in downtown New York City, to the RCA satellite uplink in Vernon Valley, New Jersey.[10]

At Vernon Valley the 384 Kbit groups are demultiplexed from the T1 format and reconfigured into the ADDS format without undergoing demodulation. Similar arrangements will be utilized by ABC and CBS. At the individual radio station the digital receiver will select the desired channel, decode it and provide a normal audio output in the radio station control room. Each affiliate will be provided with decoding modules for their respective network. The result will be to provide a virtually transparent path between the networks' program origination facilities and the affiliate station with a total of only one modulation section.

Meanwhile, common carrier operators with customers which require origination from more than one uplink on a regular basis prefer multiple access digital audio systems. AT&T plans to offer such a service based on the M/AESTRO digital audio system manufactured for AT&T by Digital Communications Corporation, a subsidiary of M/A Com (hence the name M/AESTRO).[11]

M/AESTRO is based on the Bell System's T1 digital radio which is a 1.544 Mbit system.* AT&T plans to make the digital service available in various combinations of 15 kHz and 7.5 kHz audio and data configured to make a typical package equivalent of 4-15 kHz audio channels plus 8 kHz of data and framing codes. The goal of this approach is to provide multiple channel high quality audio from small earth terminals (3 meters typical) in the presence of severe terrestrial interference.

M/AESTRO is also compatible with T1 carrier systems which will enable a terrestrial circuit to be tied directly into the satellite digital audio system decoding circuits with a minimum of processing. Like the ADDS, M/AESTRO digitizes 15 kHz audio at a 32 Kbit sample rate and 15 bit quantization which is compandered to 11 bits and raised to 12 bits including parity. Up to four 384 Kbit channels are combined to produced a 1.536 Mbit stream to which framing codes are added to procduce the 1.544 Mbit T1 data format. The T1 format is extremely common in the U.S. with literally thousands of T1 circuits carrying voice and data in use all over the country. Repeaters are low cost and reliable and can be used on 22 AWG twisted pair wire at intervals of about 1 mile.

---

*T1 can be made up from 32 Kbit sample rate x 12 bits/sample = 384 kBit x 4 channels = 1.536 Mbit + 8 Kbit data and framing for a total of 1.544 Mbits.

For satellite transmission however, and especially for small earth terminals, the data stream must be made more rugged to withstand relatively high BER caused by interference and low receiving signal-to-noise ratios. To do this, M/AESTRO takes the 1.544 Mbit stream and applies a 1/2 rate FEC (with "soft decision" sequential decoding) to reduce errors from 10^{-3} to 10^{-7} which enables decoders to operate with a signal-to-noise of less than 5 dB. Soft decision decoding, while adding more complexity to the decoder, helps improve the system's performance by masking errors more effectively than hard decision decoding. The 1/2 rate FEC effectively doubles the bit rate to 3.088 Mbit and AT&T will transmit M/AESTRO as a quadraphase shift keyed (QPSK) signal with an occupied bandwidth of 3.088 MHz (QPSK = 1 bit/Hz).

In order to reduce the potential of terrestrial interference at +10 MHz of the satellite signal center carrier and intermodulation products introduced by the several M/AESTRO carriers, AT&T will strategically locate the individual 3.088 MHz carriers approximately as shown in Figure 5. By using separate carriers per channel group (SCPC) AT&T will be able to simultaneously originate up to 5 different carriers, each with four 15 kHz audio channels (or other combinations of audio and data), simultaneously from up to 5 different uplinks. This provides the multiple access capability needed by common carriers and others requiring multiple uplink capability.

The integrity of the original audio signal can be preserved by the M/AESTRO digital audio service by employing a A/D converter at the origination point (studio), using a T1 carrier to the uplink, reconfiguring the T1 into the M/AESTRO format for transmission to the satellite, and reversing the process at the receiving end (radio station). The very robust 1/2 rate FEC coding insures that virtually error free decoding can be obtained under rather severe receiving conditions. The "soft decision" decoding process reduces the perceptibility of errors by an additional order of magnitude. QPSK saves bandwidth by encoding at a rate of only 1 bit/Hz, and the relatively narrow bandwidth of the M/AESTRO carrier at 3.088 MHz compared to the ADDS system of 15.55 MHz makes it easier to design RF circuits with adequate performance for the QPSK signal. The better protection from errors provided by the 1/2 rate FEC is needed because lower satellite power must be used for five M/AESTRO channels. Each system (DATE, ADDS, M/AESTRO) contains merits, compromises and drawbacks, but they have been carefully tailored to the services required by the satellite operator and their customers.

In the digital audio systems described above, it became clear that two were customized for a single uplink location (DATE and ADDS), with up to nineteen 15 kHz channels while the other is designed for multiple source transmission (M/AESTRO) with up to 5 separate sources possible with up to four 5 kHz audio channels in each. Further breakdown into single or dual 15 kHz digital audio channels becomes inefficient in terms of spectrum utilization and expensive due to the lack of economics associated with multiple-use digital systems. A single digital carrier for one 15 kHz audio channel occupies at least 384 kHz of spectrum and more if FEC is applied. Until such time as a TDMA system can be developed at a reasonable cost, the only true multiple access system for audio is Frequency Division Multiple Access (FDMA), which can be accomplished economically with analog FM SCPC systems, and is in widespread use today.

It is worthwhile considering current analog audio transmission techniques in this paper on digital audio. There are currently over 20 different users of analog audio circuits on satellite, each with up to 12 individual carriers (one for each audio channel) and employing a variety of satellite powers and bandwidth.[12] One example of SCPC use on a relatively large scale is the National Public Radio (NPR) system which is based in Washington, D.C.

NPR currently employs 12 SCPC circuits on transponder 2 on the Westar IV satellite. The transponder is also shared with 4 carriers of the Mutual Broadcasting System* and 2 for Muzak (now originating from the NPR facilities). Nominal bandwidth for each carrier is 200 kHz. When properly spaced in the transponder to reduce interference due to intermodulation products, cochannel signals from adjacent satellites and terrestrial microwave sources at $\pm$ 10 Mhz (from center frequency), not only do these 18 audio carriers exist now (along with some data channels), but there is room for several more. Figure 6 shows the spectrum arrangement for the SCPC transponder. However, with so many carriers, each provides a bearer channel S/N only about 47 dB at such low power levels. In order to reduce the signal-to-noise to acceptable levels (better than 70 dB), sophisticated audio companding systems must be employed with S/N improvements of up to 29 dB possible with negligible aural perception of its operation. For NPR this is accomplished with a single band, 3:1 ratio, dynamic preemphasis, and dynamically controlled attack and release time compander made by DBX.[13] It is used by both NPR and Mutual. Mutual runs higher satellite power for its 4 carriers (19 dBW), compared to 16.5 dBW for NPR, so that only 2-3 meter antennas are needed at receiving sites compared to 3-4.5 meter antennas for NPR. NPR also operates 18 uplink facilities located throughout the country (including Alaska) and the analog SCPC system permits virtually an almost unlimited combination of uplink configurations (multiple access). Muzak operates with two 15 kHz carriers

---

*Mutual plans to add six more channel on transponder 1 in the future.

at a power of 22 dBW and as a result, can use receiving antennas as small as 1.5 meters in diameter, to produce an audio signal to noise of just over 50 dB with no companding.

The economies of scale become apparent here, because higher transponder or satellite power costs more than lower power. NPR leases many low power carriers, but must use relatively larger 3-4.5 meter diameter receiving antennas. But NPR receiving sites number less than 300 compared to over 900 2-3 meter diameter antennas for Mutual and potentially thousands of 1.5 meter antennas for Muzak.

NPR carriers at 16.5 dBW typically produce an uncompandered audio signal-to-noise ratio of about 47 dB. With 29 dB of companding, the subjective S/N is raised to 76 dB. Amplitude vs. frequency response is excellent to beyond 15 kHz and distortion levels of less than 0.1% THD are typical. The result is a subjectively high quality signal.

NPR prepares the baseband of RF signals at the NPR studio facilities where individual audio channels are modulated at 70 MHz IF, transmitted via microwave to the PBS/NPR uplink and upconverted to 6 GHz for transmission to the satellite. At the receiving end, the signal is downconverted to 70 MHz and demodulated in the station's master control room. Only one modulation section is employed.

In sum, the DATE system occupies a relatively small amount of spectrum per 15 kHz audio channel (1.79 x 4 = 450 kHz), but without substantial error correction circuits, must operate with a relatively large carrier-to-noise ratio which is available by riding as a subcarrier on a video signal. The ADDS system is designed for the national radio networks to provide a large number of 15 kHz audio channels on a single transponder with a low enough bandwidth to reduce the effects of terrestrial interference and with enough error correction to permit using small diameter earth terminals with relatively low carrier-to-noise levels. The M/AESTRO system for common carriers permits up to 5 groups of 4-15 kHz audio channels to originate from up to 5 different uplinks (locations), provides multiple access capability, relatively high immunity from terrestrial interference, and heavy error correction coding which, when combined, allow the customer to use small diameter earth terminals. Finally, the analog SCPC approach to high quality audio transmission at this time, occupies the least amount of spectrum (200 kHz) and offers the ultimate flexibility in multiple access, but with heavy analog companding required to provide adequate signal-to-noise performance with small diameter earth terminals at a reasonable cost.

All of which is not to say that either analog or digital systems are better, but that it is apparent that the state of the art for both has been pushed almost to its limit and that each system has been carefully engineered to meet the requirements of the user.

References

1. "Minutes of the Meeting of the Digital Audio Technical Committee", Journal of the Audio Engineering Society, Vol. 30, No. 1/2, 1982.

2. J.E.D. Ball, "The Planning and Implementation of the Public Television Satellite Interconnection System", Journal of the SMPTE, December, 1978.

3. R. Evans Wetmore, "System Performance Objectives and Acceptance Testing of the Public Television Satellite Interconnection System", Journal of the SMPTE, December, 1978.

4. Edmund A. Williams, "Earth Terminal Site Selection and Installation for the Public Television Satellite Interconnection System", Journal of the SMPTE, December, 1978.

5. R. Evans Wetmore, "DATE: A Digital Audio System for Television", Journal of the SMPTE, March, 1974.

6. "Pulse Code Modulation for High Quality Sound-Signal Distribution: Subjective of Digital Errors", BBC Research Dept., Report No. EL-73, Serial No. 1972/40.

7. "Pulse Code Modulation for High-Quality Sound Distribution: Quantization Distortion at Very Low Signal Levels", BBC Research Dept., Report EL-40, Serial 1970/18.

8. Steven B. Salamoff, "The DATE System: Program Audio Via Existing Video Transmission Facilities", Journal of the SMPTE, June, 1979.

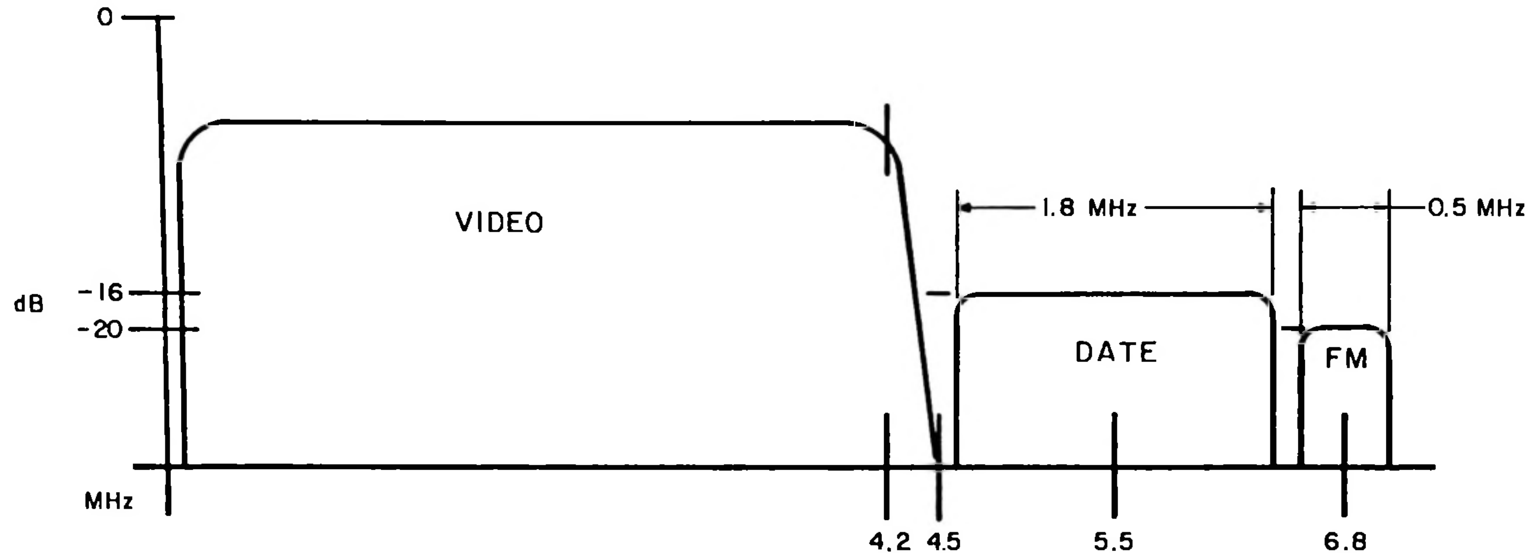
9. Ibid reference 5.

10. K.E. Fultz and D.B. Penick, "The T1 Carrier System", Bell System Technical Journal, September, 1965.

11. B.D. Bucshman, "M/AESTRO: Digital Program Audio Via Satellite", Digital Communications Corporation, DCC-0407-682-PP-1695, June, 1982.

12. "Network Notebook", (a list of network radio users on satellite), Radio World, August/September, 1982.

13. D.B. Talbot, "A Satellite Communications, Broadcast-Quality Amplitude Compander", Journal of the AES, October, 1981.



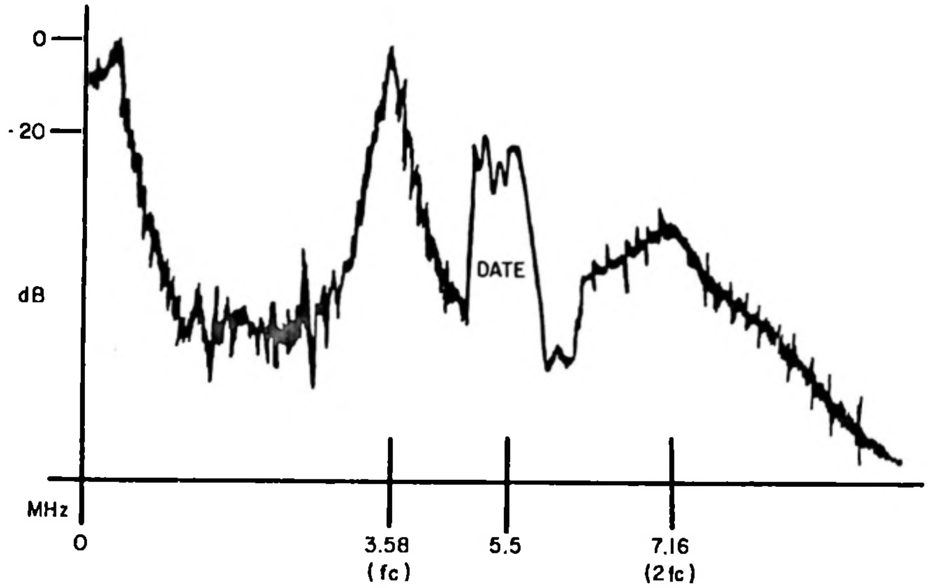
Video Baseband with DATE and FM Subcarriers

FIGURE I

TABLE I

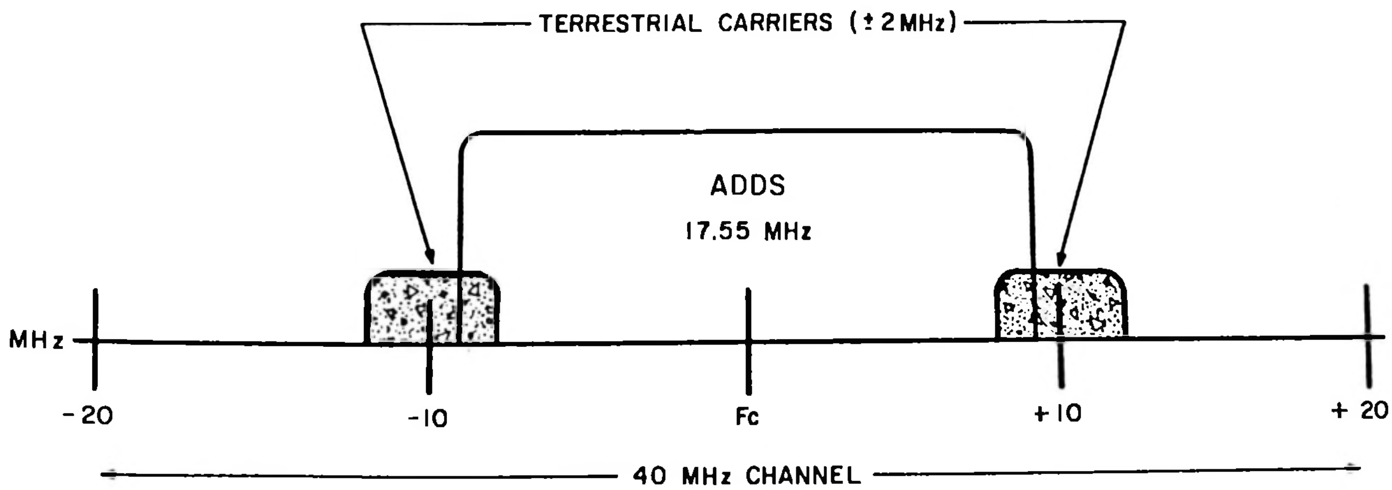
DATE PERFORMANCE PARAMETERS

- 4 - 15 kHz AUDIO CHANNELS
- ± 0.5 dB 20 Hz TO 15 kHz
- 76 dB AUDIO SIGNAL-TO-NOISE
- 0.3 % DISTORTION (THD)
- 14 BITS QUANTIZING COMPANDED TO 11 BITS
- 13 BITS TRANSMITTED (11 + PARITY + LAW)
- 34.42 kHz SAMPLING RATE
- QPSK MODULATION
- TOTAL MULTIPLEXED BIT RATE = 1.79 MBIT
  ($13 \times 34.42 \times 10^3 \times 4$ = 1.79 MBIT)
- OCCUPIED BANDWIDTH = 1.79 MHz
- OUTPUT CARRIER FREQUENCY = 5.5 MHz
- REQUIRED CARRIER-TO-NOISE RATIO = 10.5 dB



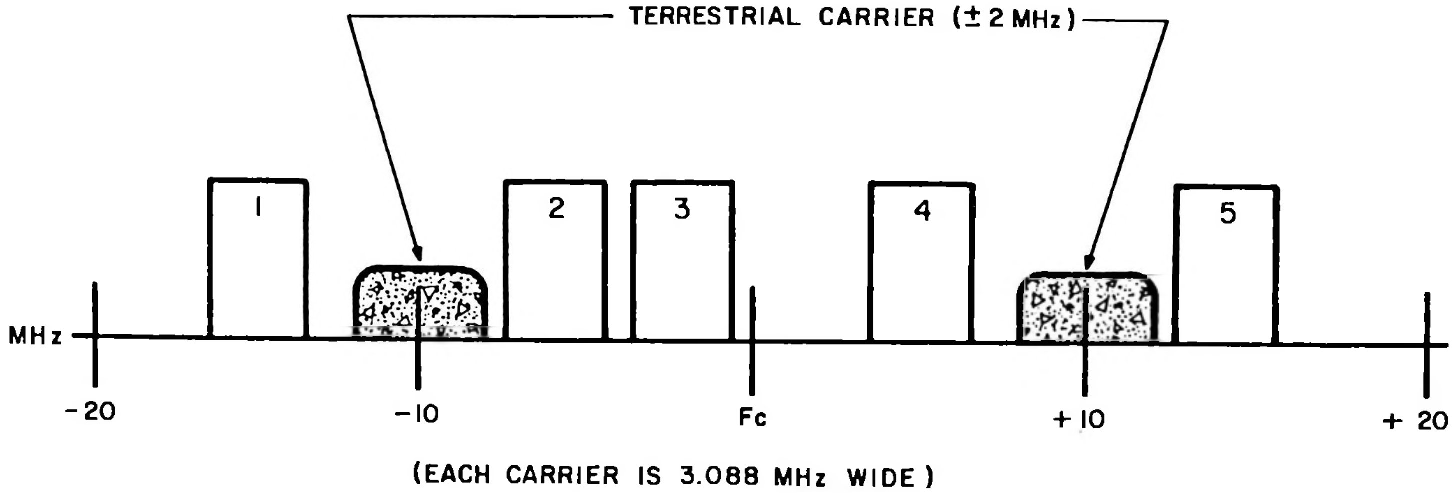
Representation of Waveform of AT & T TD Microwave System with DATE Signal added at 5.5 MHz

FIGURE 2



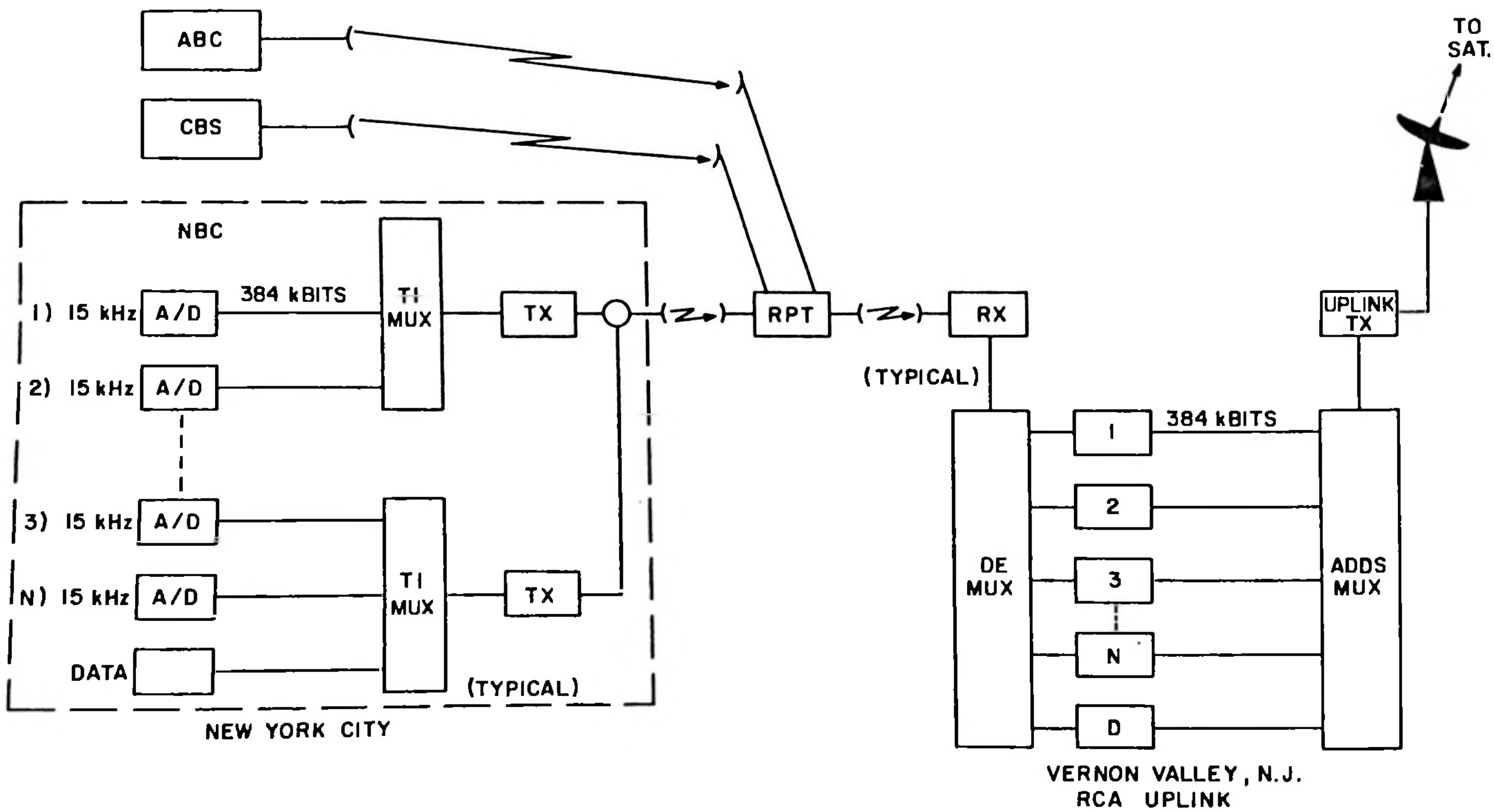
Transponder Spectrum of ADDS System

FIGURE 3



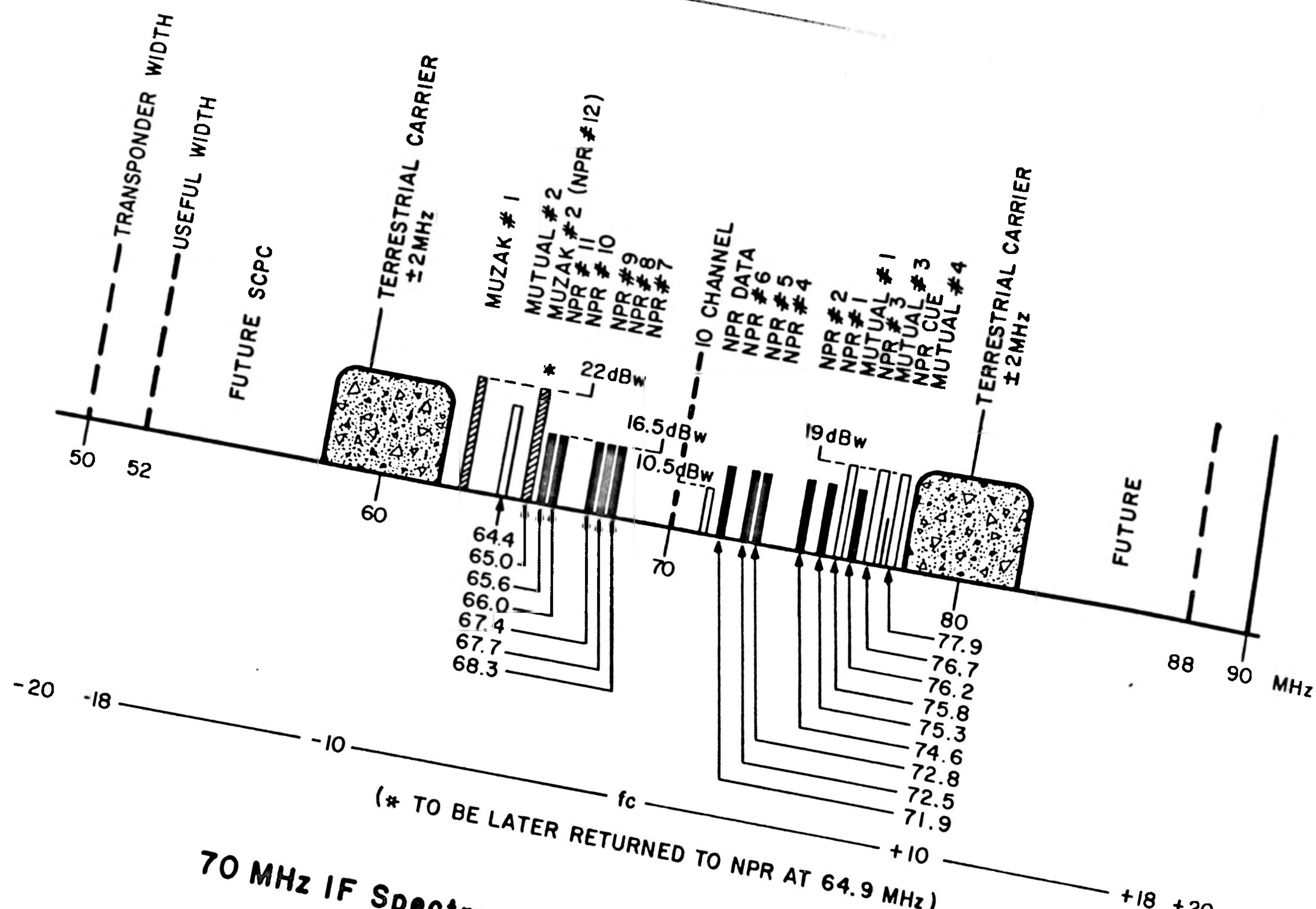
Transponder Spectrum for M/AESTRO System

FIGURE 5



Typical Routing for Radio Network Digital Audio

FIGURE 4



70 MHz IF Spectrum for NPR and Other SCPC Users

FIGURE 6

LOW POWER APPLICATIONS OF SCPC TECHNOLOGY

William A. Check

Mutual Broadcasting System

Arlington, Virginia

Abstract

There is currently a high level of interest by many broadcasters in the status and future of the regional and state radio networks--a rapidly growing population who are taking advantage of low power SCPC satellite technology.

This paper will review the technological advances which have opened new opportunities for low power SCPC applications and how they relate to the regional and state networks. It will report on the activities of the networks who are converting to satellite technology, and discuss future possibilities for these systems.

Overview

SCPC (single channel per carrier) technology became attractive to radio networks in the late 1970's when a system was developed for transmission of high quality audio material into small aperture earth terminals. Although variations on the basic design were implemented, they all utilize the same general principles: wide-band FM modulation with a high quality companding system.

With the development of the compander specifically for 15kHz audio satellite transmissions, earth terminals as small as four feet in diameter could be realized. By optimization of system parameters, earth terminals now in common operation use antennas between six and fifteen feet in diameter.

Since the late 1970's, thousands of earth terminals utilizing this type of technology have been installed. The industry has also seen several new manufacturers introduce new receiving equipment for the radio networks. As a result of this increased competition, new and innovative products have entered the marketplace which have had a major impact on systems, cost and availability.

Recently introduced into the marketplace has been a low cost narrowband demodulator. By using this demodulator and a 12-foot antenna, satellite downlink power requirements are greatly reduced. This results in lower cost for the space segment and a lower cost uplink facility.

THE NARROWBAND DEMODULATOR

National radio networks utilizing analog SCPC satellite distribution use a low cost FM wide-band demodulator developed specifically for this purpose. Most systems use a 200kHz noise bandwidth. By modification of several parameters, such as top modulating frequency, deviation, and demodulator noise bandwidth, a significant reduction can be made in the amount of bandwidth and power that are required from the satellite. This translates into a major reduction in operating costs.

During the past year, several companies have developed narrowband demodulators which have an approximate noise bandwidth of 50kHz. Some demodulators of this type have very small channel spacing, and require a high stability pilot carrier. However, by increasing the separation between channels, a cost savings can be realized. Typical requirements for channel spacing for these new demodulators used by regional and state radio networks are 200kHz.

The demodulator has two IF sections. The first section is wideband, approximately 300 kHz. The demodulator uses this wideband window to scan the spectrum and lock onto the carrier within it. The second IF section is tuned to the 50kHz demodulator noise bandwidth. AFC is included in this section, to maintain lock on the channel and to compensate for frequency errors such as minor downconverter shifting.

As with wideband demodulators, the narrowband demodulators utilize emphasis for additional signal-to-noise improvement, as well as a compander. Systems presently installed utilize a 7.5kHz audio channel. The signal-to-noise ratio for these systems is in excess of 60 dB.

SPACE SEGMENT

By using a narrowband demodulator and a 12-foot diameter antenna, a system can be configured that will minimize the satellite downlink power required. Present configurations require only 10dBw downlink power, as opposed to 19 or 22 dBw used by some other radio networks.

Western Union has made available spare transponder space for 10dBw service on both WESTAR III and WESTAR IV. Under the new Western Union tariff, the cost savings is dramatic:

19 dBw downlink power, 200 kHz channel	$11,396/month
10 dBw downlink power, 50 kHz channel	1,850/month

This substantial reduction in cost has permitted many regional and state radio networks to distribute material via satellite. By doing so, these networks have been able to take advantage of the features of SCPC transmission--multiple channels and remote uplink capabilities.

LOW POWER UPLINK

Due to the small amount of downlink power required, the corresponding uplink power required is reduced. The uplink power required is so small that equipment costs can be greatly reduced. By using an antenna as small as 4.6m (15 feet), a solid state 10 watt amplifier can be used. Using solid state amplifiers or TWT's in place of HPA's can reduce the cost as much as $80,000. Currently, low level audio uplinks can be purchased for a little as $50,000.

Much of this reduction in cost is due to competition between manufacturers. For example, one company is completing development of an uplink system that will use a directly modulated 6 GHz RF source in place of a modulator and upconverter. For users with a limited number of channels, this will provide a low cost method for uplinking service.

Using low level SCPC carriers can reduce the size of the uplink antenna required. Although some systems, due to particular requirements, may use 7m (23 feet) antennas or larger, the uplink station can be configured with antennas as small as 4.6m (15 feet).

REGIONAL AND STATE RADIO NETWORKS

Because of the low cost associated with a narrow bandwidth, low level SCPC system, many state and regional radio networks have turned to this type of technology. In addition to the low cost, it offers several advantages for uplink redundancy and system interconnection.

Several state networks are presently in operation; more are expected to become operational in the near future. Although there are some variations among state networks in the way they configure their equipment, the basic design is the same: a central low level uplink site feeding several channels on the satellite transponder. The receive terminals are equipped to receive multiple transmissions, usually several channels. The number of downlink terminals in a system is typically between 70 and 150.

Unique uplink sharing arrangements have enabled state networks to take advantage of SCPC technology. Because SCPC carriers within a transponder may be uplinked from different sites, some state networks have entered into agreements with other nearby state networks to provide emergency back-up restoration. If a catastrophic failure occurs at either station, service can be quickly restored. Several networks have already entered into these types of agreements. The Virginia Network and the North Carolina Network, with uplink sites in Richmond and Raleigh respectively, will have the capacity to provide back-up restoration for each other. The Louisiana Network and the Mississippi Network can back each other up as well. Their uplink sites are in Baton Rouge and Jackson.

Western Union has made available 10 dBw satellite channels on both WESTAR III and WESTAR IV. Each network must look at its individual needs when making a determination as to which satellite is preferable. Each satellite has its particular advantages.

Several state networks presently in operation have taken service in WESTAR III, Transponder 2. By being on a transponder with several other radio networks, the possibility of sharing channels or programming exists. This will increase in importance as more state networks become operational within the transponder. Also, it will become more attractive as manufacturers introduce new low cost agile demodulation equipment.

Another attraction of WESTAR III is the national radio networks and wire services already located in Transponder 1. Equipment manufacturers are looking at methods of providing low cost demodulation equipment for dual transponder operation with one downconverter, so that stations may receive multiple channels from multiple transponders with a minimum of equipment.

Using multiple channels, the state networks offer programming that is otherwise difficult to obtain. Most state networks have news channels and special events channels to handle regional and local sports, and other events. The state networks now have a flexibility that was previously unattainable.

SUMMARY

Due to the development of low cost downlink equipment, the regional and state networks have been able to take advantage of satellite delivery. Use of low power SCPC channels have dramatically reduced the space segment operating costs, and have had a favorable impact on uplink costs. As the number of state networks increases, the manufacturers will introduce new low cost equipment to meet the needs of these networks.

AM STEREO SYSTEM

TRANSMISSION AND RECEPTION

Leonard R. Kahn

Kahn Communications, Inc.

Garden City, New York

The AM Stereo system proposed by Kahn Communications and the Hazeltine Corporation is an Independent Sideband System wherein the lower sideband represents the left stereo channel and the upper sideband represents the right stereo channel. Thus, this system is essentially a frequency separation system as distinguished from the other major systems which are essentially phase separation systems in that the stereo separation is achieved by detecting the phase difference of stereo components.

It is the author's belief that this distinction; i.e., frequency separation vs phase separation, is the basis for the differences in performance of the systems especially under adverse conditions such as co-channel and adjacent channel interference, multipath conditions such as skywave/ground wave interference, sensitivity to transmitting and receiving antennas, and capability of using advanced receiver technology such as Asymmetrical Sideband Selectivity.

While the means for generating an ISB wave is, for the block shown below, phase sensitive, nevertheless, once the independent sideband wave is generated stereo separation is insensitive to phase as the wave is passed through the transmitting antenna, the propagation path, the receiving antenna, the receiver's IF and RF circuits, and the input circuit feeding the stereo demodulator.

Transmitting Equipment

Please see FIG. 1 which is a simplified block diagram of the type of AM Stereo generator used in the STR-77 unit. The left and right inputs are summed and fed through a constant amplitude, phase difference network. The output of this network feeds an audio frequency amplifier which produces an output suitable for modulating conventional AM transmitters.

The difference between the L and R components is fed to a companion phase difference network which assures a quadrature relationship between the output of the sum circuit and the difference circuit. This output is then fed to a phase modulator. The resulting phase modulated wave is frequency translated to the carrier frequency of the station. An RF amplifier is used to raise the level to 3 watts, sufficient for exciting most AM transmitters. This phase modulated wave is then envelope modulated in the station's transmitter to produce an independent sideband wave wherein the lower sideband carries the left information and the upper sideband carries the right information.

In a practical configuration, circuitry is provided for insuring proper time delay so that the AM and PM components arrive at the transmitter's modulated stage coincidentally. Additional circuitry is shown in FIG. 1 for producing a second order L-R wave component which is added to the fundamental L-R audio wave fed to the phase modulator. This second order component minimizes bandwidth and allows low distortion reception.

FIG. 2 shows a simplified phasor sketch showing how the first order sideband components line up so that the lower sideband represents the left stereo information and the upper sideband represents the right stereo information. The following equation is a descriptive equation of the modulated wave:

$$e_s = E_c \left[1 + m_L\, L(t)\, e^{J\,\pi/4} + m_R\, R(t)\, e^{J\,\pi/4} \right]$$

$$\times \cos \left[w_c t + m_L\, L(t)\, e^{-J\,\pi/4} - m_R\, R(t)\, e^{-J\,\pi/4} \right.$$

$$- .532 \left[m_L\, (L(t)\,)^2 - m_R\, (R(t)\,)^2 \right]$$

$$\left. \times \frac{k}{T} \int_0^T \left| m_L\, L(t) - m_R\, R(t) \right| dt + \lambda \sin 2\pi\, 15t \right]$$

Receiver

FIG. 3 is a simplified block diagram of a receiver suitable for demodulating an independent sideband stereo wave. Details concerning the carrier track circuitry and the inverse modulation circuit is contained in the referenced patents and Hazeltine's report.*

Asymmetrical Sideband Selectivity

The ideal receiver for the instant stereo system provides enhanced selectivity for adjacent channel interference by taking advantage of the fact that adjacent channel interference is naturally separated so that lower channel interference falls solely in the left stereo channel and higher channel interference falls in the right stereo channel. This natural characteristic is the basis for the "cocktail party effect" which causes interference to fall full to the right or full to the left edge of the stereo stage according to whether the interference is above or below the desired channel.

In contrast, phase separation systems produce interference in both the left and right channels, causing the interference to fall "on stage." The natural separation of interference on one channel or the other in the present system also allows the use of a new form of selectivity which is called Asymmetrical Sideband Selectivity. (Please see FIG. 4 which is a simplified drawing of asymmetrical sideband selectivity.)**

It is noted that the interference above and below the desired channel is constantly monitored. Under low interference conditions, as would be expected when listening to the local stations or during daytime reception, the receiver's bandpass is wideband, providing high fidelity performance. When significant interference is detected, the channel with interference is automatically reduced in bandwidth, attenuating the offending adjacent channel interference.

---

*L. R. Kahn, U. S. Patent 4,018,994 dated April 19, 1977.
L. R. Kahn, U. S. Patent 3,973,203 dated August 3, 1976.
Hazeltine Research Inc., Note 1, April 5, 1982, "ISB AM Stereo Receiver Practices."

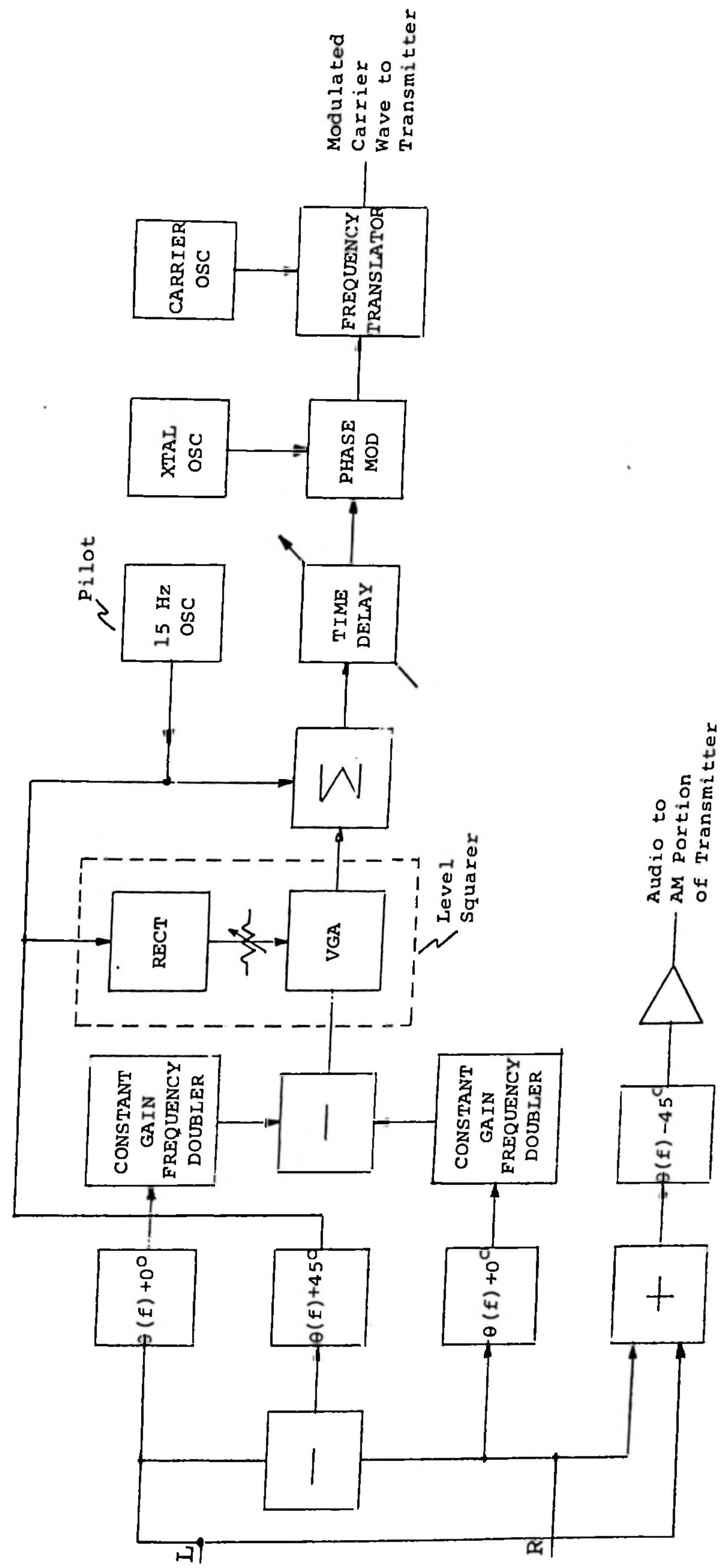
**L. R. Kahn, U. S. Patents 4,192,970 and 4,206,317

Since under normal conditions the interference is not symmetrical the frequency response of the overall system is restricted by stereo channel (L or R) having the higher response because it has the least interference. However, there is a 6 db step in the overall response because only one channel would have a 10 kHz signal level, for example. Thus, by use of the "mixed highs" concept the frequency response is degraded by much less drastic loss; i.e., 6 db rather than say, 20 db or more. This, of course, is substantially better response than would be the case if both channels were reduced in bandwidth symmetrically. It is important to note that this type of adaptive selectivity is also useable for monophonic reception of mono signals.

It is the author's opinion that this improvement in effective selectivity, since it can be implemented by use of newly developed***IC technology at very low cost, may well be the main advantage in implementing the ISB AM Stereo system and may have a major effect on the future growth of AM broadcasting.

---

* **R. W. Brodersen, P. R. Gray and D. A. Hodges, "MOS Switched-Capacitor Filters", PIEEE, vol. 67, pp. 61-75, January 1979.



Modulated Carrier Wave to Transmitter
CARRIER OSC
FREQUENCY TRANSLATOR
XTAL OSC
PHASE MOD
Pilot
15 Hz OSC
TIME DELAY
Σ
RECT
VGA
Level Squarer
CONSTANT GAIN FREQUENCY DOUBLER
CONSTANT GAIN FREQUENCY DOUBLER
Audio to AM Portion of Transmitter
θ(f) +0°
θ(f)+45°
θ(f) +0°
θ(f) −45°
+
−
−
L
R

FIG. 1

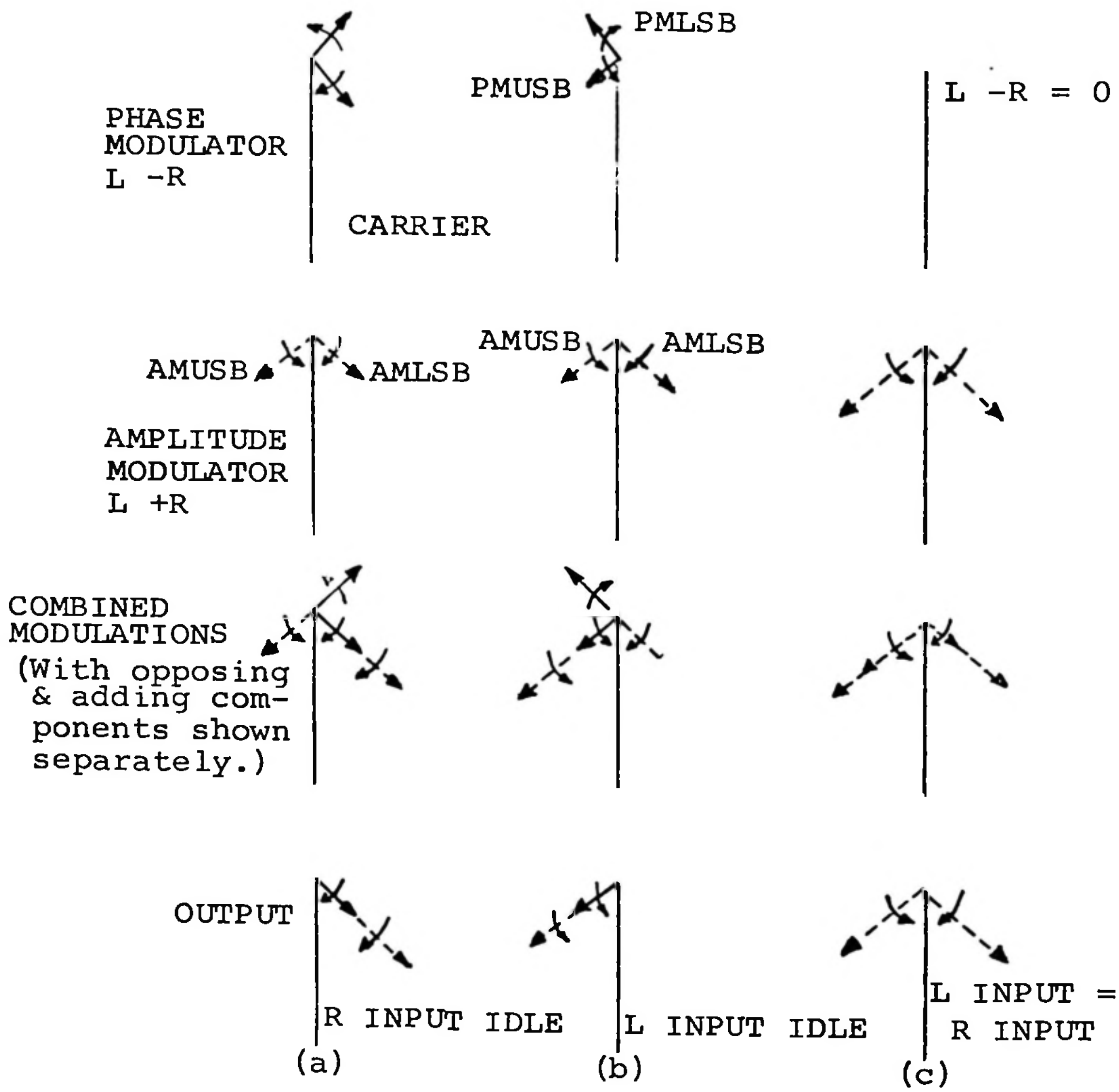

PMLSB
PMUSB
L -R = 0
PHASE
MODULATOR
L -R
CARRIER
AMUSB
AMLSB
AMUSB
AMLSB
AMPLITUDE
MODULATOR
L +R
COMBINED
MODULATIONS
(With opposing
& adding com-
ponents shown
separately.)
OUTPUT
R INPUT IDLE
L INPUT IDLE
L INPUT =
R INPUT
(a)
(b)
(c)

FIG. 2

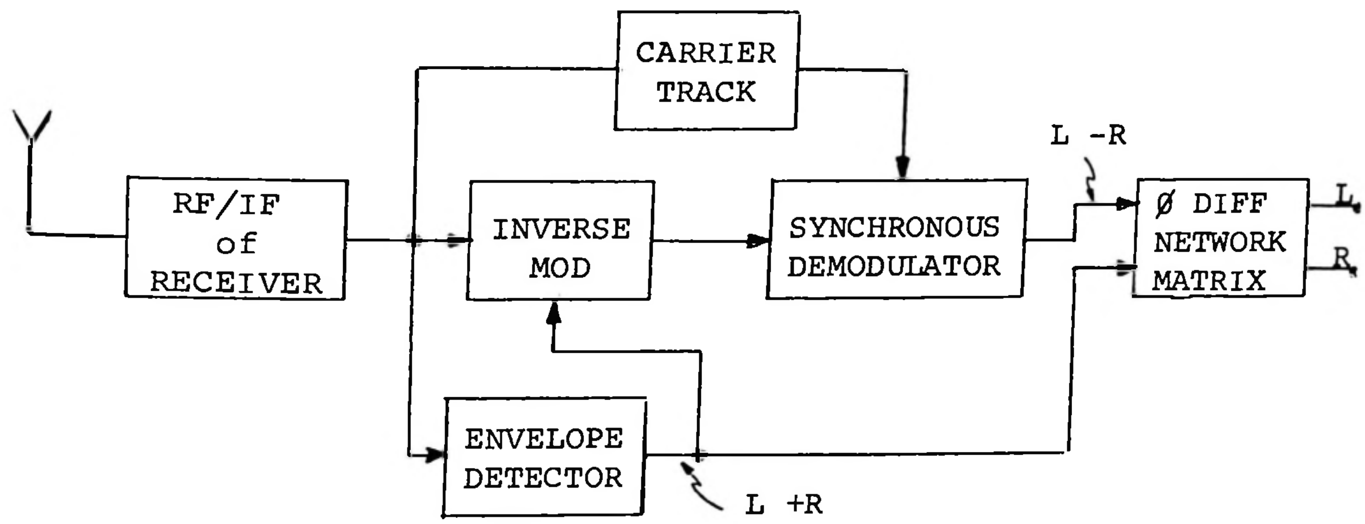

CARRIER
TRACK
RF/IF
of
RECEIVER
INVERSE
MOD
SYNCHRONOUS
DEMODULATOR
L -R
Ø DIFF
NETWORK
MATRIX
L
R
ENVELOPE
DETECTOR
L +R

FIG. 3

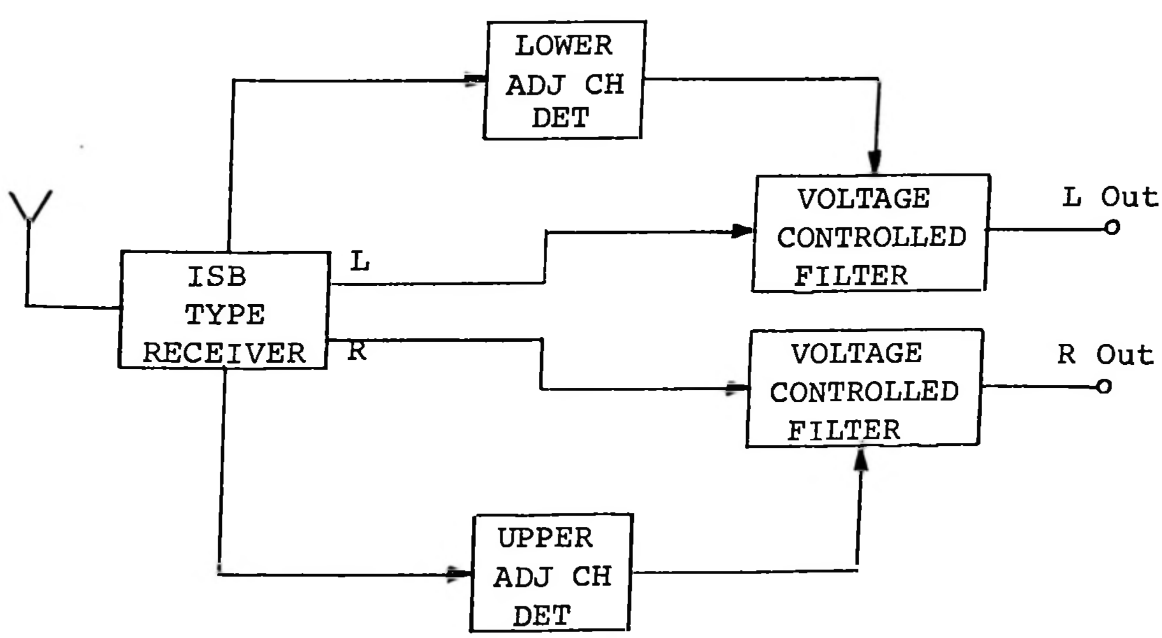

LOWER
ADJ CH
DET
VOLTAGE
CONTROLLED
FILTER
L Out
ISB
TYPE
RECEIVER
L
R
VOLTAGE
CONTROLLED
FILTER
R Out
UPPER
ADJ CH
DET

FIG. 4

THE HARRIS LINEAR AM STEREO SYSTEM

David Hershberger

Harris Corporation

Quincy, Illinois

I. Introduction

Harris promotes linear quadrature modulation and synchronous detection for AM stereo. Although envelope detectors produce some distortion when receiving linear quadrature modulation, the distortion is almost entirely second harmonic, which is seldom audible and not objectionable. The Harris AM stereo system incorporates special patent pending audio processing techniques which ensure that envelope detector distortion will not be objectionable to the listener. Major broadcasters using the Harris linear system are very pleased with both stereo performance and monophonic compatibility.

AM stereo can appear deceptively simple to the casual investigator. Although amplitude modulation is a "low-technology" mode, AM stereo, and especially the selection of an AM stereo standard, is a complex issue requiring special insight. First impressions are usually wrong in AM stereo.

A thorough investigation of AM stereo, including a fundamental evaluation of compatibility, will show that linear systems are best suited for AM stereo.

The primary reason for AM stereo is to allow AM stations to compete more effectively with FM by improving the technical quality of the AM service. Although AM does have a higher noise level than FM, the audio quality limitation is primarily due to poor radios. Most AM stations transmit audio to beyond 10 kHz, but most radios have a frequency response of only 2 to 3 kHz. Virtually all radios use the antiquated envelope (diode) detector to recover modulation. Envelope detectors are inherently prone to the production of various types of distortion and other desireable effects. AM radios can be vastly improved by extending the bandwidth for better frequency response, and be replacing the envelope detector with a synchronous detector for reduced distortion.

II. Nonlinear Systems

A linear system is one in which superposition applies; that is, the system response to several inputs applied simultaneously is the same as the sum of the system responses when the inputs are applied individually. A nonlinear system is one in which superposition does not apply.

Proponents of nonlinear modulation use it in an attempt to obtain compatibility with envelope detectors. If receiver bandwidth were infinite, there would be perfect compatibility.

One way of analyzing nonlinear modulation is as a predistorted system. Because the envelope detector is also a nonlinear device, nonlinear systems must predistort the upper and lower sidebands of an AM stereo signal in order to force the envelope to be L+R.

This is equivalent to transmitting distortion 180 degrees out of phase with the distortion the envelope detector generates. Although such a signal is perfectly compatible with envelope detectors, it is unfortunately incompatible with another circuit found in all AM radios: the bandpass filter.

It can be shown that nonlinear systems have an infinite number of sidebands, and in practice three or four orders of these sidebands are significant. The high order or predistortion sidebands must all be present at the receiver's detector for proper operation and compatibility. But by the time a nonlinear AM stereo signal passes through the IF bandpass filter of a typical narrow bandwidth radio, the relative amplitude and phase relationships of the required predistortion sidebands may be altered such that distortion is unacceptable, even though the radio may be correctly center-tuned. Truncation of the spectrum by the bandpass filter's amplitude characteristic, and phase distortion caused by the filter's group delay characteristic, will cause the predistortion sidebands to arrive at the wrong amplitude and at the wrong time to "cancel" the envelope detector distortion.

There are two significant types of compatibility distortion with nonlinear systems: harmonic distortion and difference tone intermodulation distortion. Of the two, the difference tone distortion is more objectionable to listeners.

It may be argued that these types of distortion are of little significance since program material generally contains little high frequency energy. That argument, however, is invalid because of the general use of pre-emphasis on AM. To partially compensate for the limited frequency response of AM radios, most AM broadcasters boost the higher modulating frequencies. Since the improvement in received signal quality can be quite dramatic, broadcasters will be unwilling to sacrifice the use of pre-emphasis.

Pre-emphasis, which frequently boosts the higher frequencies as much as 20 to 30 dB, severely aggravates the filtering-related compatibility problem of nonlinear signals.

III. Linear Systems

Linear systems are those for which superposition applies. Two important characteristics of linear systems are:

1. Only one set of sidebands is produced - bandwidth is the same as mono.

2. Linear product detection in both L+R and L-R channels means that there is no "noise burst" problem whatever.

Linear quadrature modulation is simply expressed as:

$$S(t) = [1 + L(t) + R(t)]\sin(wt) + [L(t) - R(t)]\cos(wt) \qquad \text{EQ. (1)}$$

where:

$S(t)$ = stereo AM signal
$L(t)$ = left channel modulating signal
$R(t)$ = right channel modulating signal
w = carrier (radian) frequency
t = time

The stereo signal consists of a monophonic sum signal (the sine term) with the linear addition of a double sideband suppressed carrier left minus right signal in quadrature with the carrier (the cosine term). Since both signals (sine and cosine terms) have only one set of sidebands, and since addition is a linear process, the resulting stereo signal also has only one set of sidebands - no new frequencies are created by adding the two signal components.

Linear quadrature modulation is highly desireable for AM stereo broadcast application because of its many advantages over nonlinear systems.

1. Same RF bandwidth as mono
2. No distortion from spectrum truncation
3. Compatible with pre-emphasis
4. Compatibile with synchronous detection
5. Compatible with envelope detection (with audio processing)

Synchronous detectors, in turn offer the following advantages over envelope detectors:

1. No distortion due to receiver mistuning
2. No distortion due to selective fading or AM multipath
3. No distortion from directional transmitting antennas (far-field overmodulation)
4. No distortion from modulation overshoots
5. No distortion due to co-channel interference
6. No inherent cross mdoulation
7. Some reduction of impulse noise.

IV. Compatibility with Human Hearing

Compatibility is without doubt the most complex AM stereo requirement. To determine the extent to which a system is compatible, it is necessary to evaluate the end result, which is the listenability of a system on a narrow-band envelope detector radio, transmitting real program material to real human beings. The final test of compatibility comes when the radio station transmitting AM stereo goes into its ratings period.

Other tests, such as transmitting tones through wideband envelope detectors to instruments, while valuable in the quantitative sense, are not necessarily measurements of compatibility with human hearing.

It is in the determination of compatibility that errors in system evaluation are most likely to be made. If compatibility is not thoroughly investigated, it is very likely that one could conclude that a compatible system should be rejected in favor of another system that is actually incompatible with existing radios.

When defining compatibility, one's first impulse would be to state that a compatible system should have low distortion. Distortion usually is taken to mean total harmonic distortion or THD. This is the pitfall to be avioded - a simplistic view of distortion.

The point is this: the criteria for the evaluation of distortion should accurately represent the sensitivities of the human ear. For example, second harmonic distortion is only objectionable when very large. On the other hand, difference tone intermodulation distortion is quite objectionable to the ear and should weigh heavily in any system evaluation.

Linear systems have excellent potential in every area except one: there is an apparent lack of compatibility with envelope detectors. The envelope of eq. (1) is:

$$e(t) = \sqrt{[1+L(t)+R(t)]^2 + [L(t)-R(t)]^2}$$

Which is a distorted version of the desired signal, $1+L(t)+R(t)$. If a simple total harmonic distortion (THD) calculation or measurement were made, the value would appear to be too high. For example, envelope THD values for several single channel modulation levels are given in Table 1.

Left Channel Modulation	THD
75%	28.9%
50	16.1
25	6.7

Table 1. Envelope distortion vs. single channel modulation

But what simple THD values do not convey is that the distortion is primarily second harmonic, which makes the distortion relatively innocuous.

Second harmonic is the least objectionable of the harmonics. In general, even-order harmonics, such as second, fourth, sixth, etc., are less objectionable than odd-order harmonics such as third, fifth, seventh, etc. Furthermore, the lower the order of the harmonic, the less objectionable and audible it is. Second harmonic is much less audible and much less objectionable than is, say eleventh harmonic.

These general characteristics of human hearing are in turn due to the physiology of the ear/brain combination, and how we hear sounds. The ear, for instance, has a large amount of second harmonic built in. If two flutes are played at moderate volume, the ear will create a third tone due to the difference in frequencies between the two flutes. In fact, some pieces of music take advantage of the phantom difference tone created by the ear. Several Telemann flute sonatas create a difference tone part which harmonizes with the "real" parts. A more familiar example of music with a third part created by the ear is "Yankee Doodle".

In fact, there is some indication that second harmonic distortion is in some instances actually desireable. One example is a popular piece of studio equipment which is essentially a second harmonic generator. This unit is widely used by recording studios to process music for increased appeal.

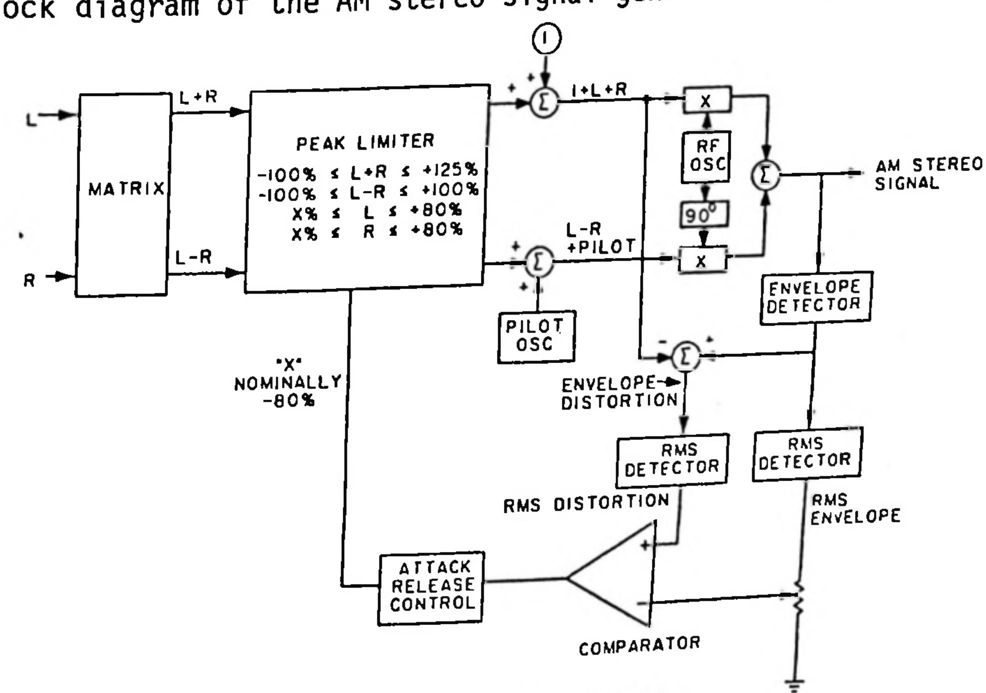
The addition of second harmonic, instead of sounding "distorted", sounds like increased treble response. While conventional "equalization" or high frequency boost also increases the noise, artificial highs from a second harmonic generator will not boost the noise. In summary, adding second harmonic sounds basically like treble boost, but without a background hiss boost.

V. Controlled - Compatibility Linear Quadrature

It has been the finding of Harris and other investigators that linear quadrature modulation is "almost compatible". If something minor could be done to obtain compatibility without sacrificing the linearity of the system, then a truly superior AM stereo system would be feasible.

Harris has developed a special audio processing technique which ensures compatibility with envelope detectors, while not significantly affecting the stereo signal. The audio processing (compatibility control) works by determining the distortion which would occur in an envelope detector; when the envelope distortion exceeds an acceptable level, single channel negative modulation constraints are reduced. Normally, single channel left or right signals are allowed to modulate up to +80% and -80%. When envelope detectors produce excessive distortion, the negative modulation constraint for single channel signals is reduced from 80% to something less; typically -70% or -60%. The negative constraint is reduced until the envelope detector distortion is within acceptable limits.

A block diagram of the AM stereo signal generator is shown in Figure 1.



SYSTEM CONFIGURATION FOR COMPATIBILITY CONTROL. MODULATION CONSTRAINTS ARE ADJUSTED AS A FUNCTION OF PROGRAM MATERIAL TO MAINTAIN COMPATIBILITY WHILE TRANSMITTING LINEAR QUADRATURE.

Figure 1

The time constant associated with the compatibility control processor is somewhat slow, in the hundreds of milliseconds. The deliberately slow time constant reflects the design philosophy of having a system which is designed for the human ear rather than for the sake of instrumentation. In order for the ear to detect low orders of distortion (such as second order), the distortion must be present for some time. Unlike crossover distortion, for example, which can be detected immediately, quadrature envelope distortion must be present for several hundred milliseconds before it is detectable. Accordingly, the audio processing, whose purpose it is to model the ear, is made to have a similarly slow time constant.

Broadcasters using the Harris linear quadrature system for AM stereo are unanimous in finding both excellent stereo performance and good quality compatible reception on monophonic radios. The "bottom-line" proof of the system's compatibility lies in the ability to satisfy program directors, and in station market share ratings.

VI. Conclusion

Linear quadrature modulation is very appealing for application in AM stereo. Many early attempts at using quadrature modulation compromised the benefits of the system. Some systems simply reduced the amplitude of the L-R channel, thereby incurring a signal-to-noise ratio penalty. Other systems sacrificed the linearity of the system in various ways, forcing the transmitted envelope to be 1+L+R. These nonlinear systems occupy excessive bandwidth, and in spite of the fact that the transmitted envelope is 1+L+R, radio IF bandpass filtering causes the nonlinear system to generate distortion at the radio's envelope detector. Nonlinear systems, in the final analysis, have serious compatibility problems.

The supposed "incompatibility" of linear quadrature modulation is not a real problem. When human beings are used to determine compatibility, minor audio processing makes the use of linear quadrature modulation an ideal solution to the AM stereo problem.

Broadcasters using the Harris linear system are pleased with both stereo and mono performance, while broadcasters using nonlinear systems often voice dissatisfaction with monophonic compatibility, especially when pre-emphasis is used.

The use of linear quadrature modulation will encourage the development and marketing of long-overdue improvements to the AM sections of radios: synchronous detection and wider IF bandwidths, for less distortion and wider frequency response. These improvements and the use of linear quadrature modulation are most consistent with the goals of the AM broadcaster and the consumer: to make AM stereo sound competitive with FM stereo.

TECHNICAL PERFORMANCE OF THE

MAGNAVOX PMX AM STEREO SYSTEM

J. S. SELLMEYER, P. E.

N.A.P. CONSUMER ELECTRONICS CORP.

KNOXVILLE, TENNESSEE

INTRODUCTION

The Magnavox PMX AM Stereo system uses a combination of conventional envelope and linear phase modulation to transmit a compatible stereophonic broadcast signal within the spectrum constraints of a standard medium wave broadcast channel. The vector sum of the left and right audio channels is transmitted as envelope modulation on a carrier which is simultaneously phase modulated with the vector difference of the left and right audio channels. The peak phase deviation of the difference channel of one radian corresponds to 100% envelope modulation of the carrier. Delay equalization is employed as required to ensure that peak phase deviation corresponds precisely to peak envelope modulation. This relationship provides best separation and allows simple calibration references to be used for verification of proper operation.

The mathematical representation of the PMX AM Stereo signal is as follows:

$$\left[1+m\Big(L(t)+R(t)\Big)\right] \cos \left\{(W_c+A\cos W_o t)t+B\Big(L(t)-R(t)\Big)\right\}$$

Where
- m = AM modulation index for sum channel
- B = PM modulation index for difference channel
- A = FM modulation index for identification tone
- $W_o = 2\pi$ (5) hz
- $W_c = 2\pi$ (carrier frequency) hz

A block diagram of the transmitting system is shown in Figure 1. A block diagram of a typical receiver is shown in Figure 2.

The PMX System has the advantages of simple implementation and operation, and is easily understood both from a transmission and reception hardware standpoint. There is no requirement for complicated circuitry to control the gain of the L-R channel to maintain envelope compatibility. Nor is there any requirement for exotic sideband cancellation circuitry or ANY form of linearity compensation or predistortion circuitry. There are no artificial limits or other constraints placed on the transmission system. The only limit is that of negative overmodulation. The carrier must not be permitted to be cut off for long periods of time. The difference channel information cannot be recovered during this interval, since it is not being transmitted, hence stereophonic definition will be lost. This constraint is equally true for all systems and is not unique to the PMX system. The PMX system has no other constraints on the envelope or the audio channels.

MEASUREMENT METHODS AND TECHNIQUES

Many measurement methods and techniques common to monaural AM may be applied to PMX AM Stereo as well. A little thought about all possible transmission and reception paths and conditions will reveal the significant tests which should be run. For instance, monaural compatibility with envelope detectors is an important consideration, at least for the next few years. A signal, either monaural or stereophonic may reach the monaural listener by several paths. The vector sum of the left and right channels in phase with equal amplitudes, (L+R) for example, comes to mind immediately as the most obvious. This is a common test for all AM Stereo systems and is frequently labeled as a "monaural compatibility test." It is obvious by inspection of the block diagrams of all AM Stereo systems that there should be no difference between the systems operating under these conditions since they should all produce only envelope modulation under this condition. However, a little thought will quickly reveal that the monaural listener will also hear as monaural material any non-correlated single channel information which is transmitted.

Material transmitted predominately or exclusively on either the left channel or the right channel will be perceived as monaural material by the monaural listener. In the PMX system, this material may be permitted to fully modulate both the envelope and phase channels, thus providing a better stereophonic signal to noise ratio. The PMX system is the only system with no artificial single channel envelope modulation constraints. A measurement of harmonic and intermodulation distortion using an envelope detector measuring a high level left only or right only signal would be an appropriate test for monaural compatibility. The PMX system will exhibit distortion measurements very close to those of the monaural condition. Any minor difference is solely a function of the receiving hardware and is predominantly due to cross mode conversion from phase to amplitude modulation, or vice versa. Certain systems will exhibit alarmingly high harmonic and intermodulation distortion under this common condition.

MEASURED TRANSMITTER DATA

Data has been measured over many different radio stations in the United States as part of the NAMSRC tests and in later responses to Federal Communications Commission Rulemaking Proceedings. Tests have been conducted utilizing both non directional and directional antenna systems, some having null depths exceeding thirty decibels. Transmitter designs have ranged from pre World War II high level designs to the latest Pulse Width Modulated designs. In all cases stereophonic performance was limited primarily by the monaural capabilities of the transmission system.

Tests conducted in late 1980 at Station WOWO, Ft. Wayne, Indiana, revealed monaural harmonic distortion averages in the range of one half of one percent over the range of 50 hertz to five kilohertz. These numbers were virtually identical for L+R, left only and right only modulation sources. No significant difference was noted between the directional and non directional antenna measurements. Intermodulation distortion using the SMPTE method measured in the range of 3.0 percent on the non directional antenna for all conditions and in the range of 4.5 percent on the directional antenna. The higher intermodulation distortion measured on the directional antenna is probably due to differences in the RF load at the sideband frequencies between the non directional and directional antenna systems. The high frequency utilized in the SMPTE IM test is seven kilohertz. One would expect to see similar differences in the monaural high frequency performance as well. The important point to note is that no degradation in monaural performance resulted. The data was measured at 42.5 percent envelope modulation and demodulated with a standard envelope detector type of modulation monitor.

The stereophonic harmonic distortion performance measured on this system averaged on the order of one half of one percent for both left and right channels measured through a PMX stereo monitor connected to the transmitter monitor output.

The noise floor for monaural operation measured 53 dB below 42.5% envelope modulation. Adding a 7.4 dB correction factor to move the reference to 100% envelope modulation, the monaural noise floor was approximately 60.4 dB below 100% modulation. For the stereophonic case, the noise floor was measured at 51.75 dB below 42.5% envelope modulation. This corresponds to 59.15 dB below 100% envelope modulation. Thus, there is very little degradation of the noise floor due to stereophonic operation.

There was essentially no difference in measured frequency response between monaural and stereophonic operation.

The separation averaged 35.9 dB over the range from 50 Hz to 5 kHz.

OVER THE AIR DATA

Data was measured over the air at a location in Fort Wayne, Indiana, approximately five miles from the WOWO transmitter, utilizing a modified Philips High Fidelity Laboratories Model AH673-44 Tuner. This unit features a dual bandwidth IF in the AM section along with other features which make it desirable for conversion to AM Stereo use. The AM Stereo modification uses the National LM-1981 AM Stereo decoder integrated circuit along with a Magnavox developed pilot detector. It is possible to force either monaural or stereophonic operation regardless of the pilot transmission status for test purposes. The Left and Right outputs were used for stereophonic measurements and the Left output was arbitrarily chosen for monaural measurements. Data was taken at both 42.5% and 85% envelope modulation. Data was taken using a laboratory closed circuit setup and repeated with over the air measurements utilizing the WOWO transmission system in order to derive comparative measurements which would indicate any degradation due to the transmission system.

The following data relates to 42.5% envelope modulation.

The stereophonic harmonic distortion measured over the air between 50 Hz and 10 kHz averaged 1.36% for closed circuit laboratory tests and 1.37% for the over the air tests. For the 50 Hz to 5 kHz region, the laboratory closed circuits averaged 1.26% versus 1.25% for the over the air test. Thus, there is no meaningful measurable difference between the closed circuit and over the air tests and no degradation due to the transmission system.

The separation measured between 50 Hz and 5 kHz averaged 27 dB for the closed circuit tests and 28.3 dB for the over the air tests.

Frequency Response variation between laboratory and over the air signal conditions exhibited similar characteristics. The closed circuit measurements showed a variation of 0.3 dB between 50 Hz and 5 kHz while the over the air measurements showed a variation of 0.31 dB. For the case of full bandwidth of 50 Hz to 10 kHz, the closed circuit measurements showed a variation of 1.2 dB for the closed circuit measurements and 1.27 dB for the over the air measurements. Again, there is no significant degradation between the closed circuit measurements and the over the air measurements.

The noise measurements were made both with and without a high pass (400 Hz) filter to identify and separate the low frequency subaudible contribution. The noise floor in the wide band position measured approximately 41 dB below 42.5% modulation which corresponds to 48.4 dB below 100% modulation. The noise floor decreases to 58 dB below 42.5% modulation with a 400 Hz high pass filter in the circuit for the closed circuit case and 55 dB below 42.5% modulation for the over the air case. This results in a noise floor below 100% modulation of 65.4 dB for the closed circuit case and 62.4 dB for the over the air case.

For the stereophonic mode, the center channel noise floor measured 52 dB below 42.5% modulation or 59.4 dB below 100% modulation for the closed circuit case and 49.5 dB below 42.5% modulation or 56.9 dB below 100% modulation for the over the air case. The noise floor for a left only or right only signal measured 58 dB below 42.5% modulation or 65.4 dB below 100% modulation for the closed circuit case and 55 dB below 42.5% modulation or 62.4 dB below 100% modulation for the over the air case. The above figures were measured using the 400 Hz high pass filter. The increase in signal to noise ratio when utilizing a left only or right only transmission is attributed to the addition of the difference channel contribution to the output signal. The desired signals add coherently while the noise components add in a random fashion to provide a net increase in signal to noise ratio.

The only significant difference in the measurements taken at 42.5% and at 85% envelope modulation are in the area of distortion with both the closed circuit and over the air distortion magnitudes being somewhat higher when measured at 85% envelope modulation. Both transmitter and receiver distortion components increase with increased modulation percentage. Harmonic distortion increased to 1.8% for the closed circuit case and 1.9% for the over the air case. No other significant differences were noted.

SUMMARY

The information presented herein was derived from FCC filings submitted in response to FCC Rulemaking Proceedings. It can be seen from the data that little if any degradation to the present monaural service can be expected from installation of the Magnavox PMX AM Stereo System and in certain cases, such as Stereophonic Noise performance, enhancement of the present system is possible.

Equipment for the Magnavox PMX AM Stereo System is available from the following sources:

Broadcast Exciters
- Continental Electronics Mfg. Co., (Dallas, Texas)
- JNS, (Melbourne, Australia)

Broadcast Monitors
- Belar Electronics Laboratories, Inc., (Devon, Pennsylvania)

Laboratory Signal Generators
- Boonton Electronics Corp., (Parsippany, New Jersey)
- National, Matsushita Electric (Osaka, Japan)
- Meguro, (Tokyo, Japan)

FIG. 1

AM/PM SYSTEM BLOCK DIAGRAM

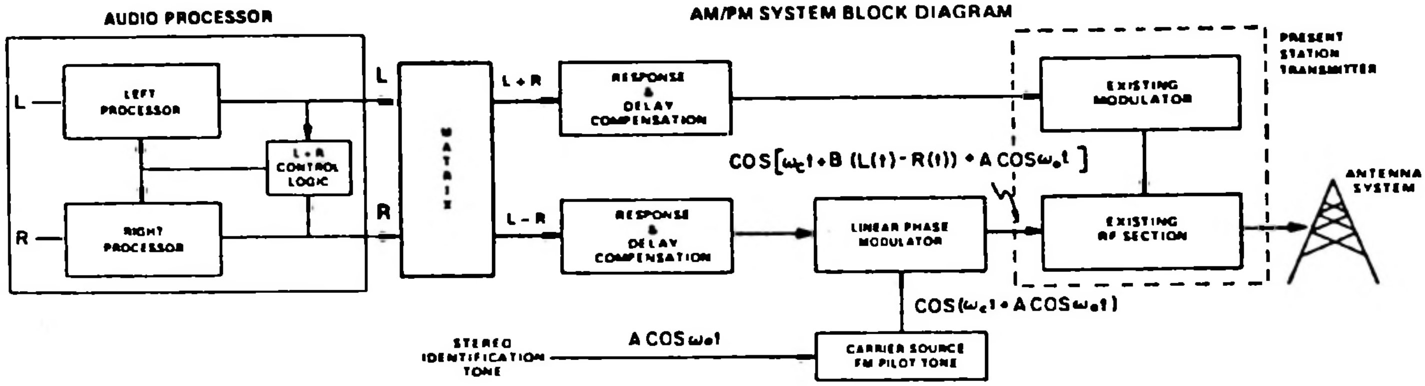
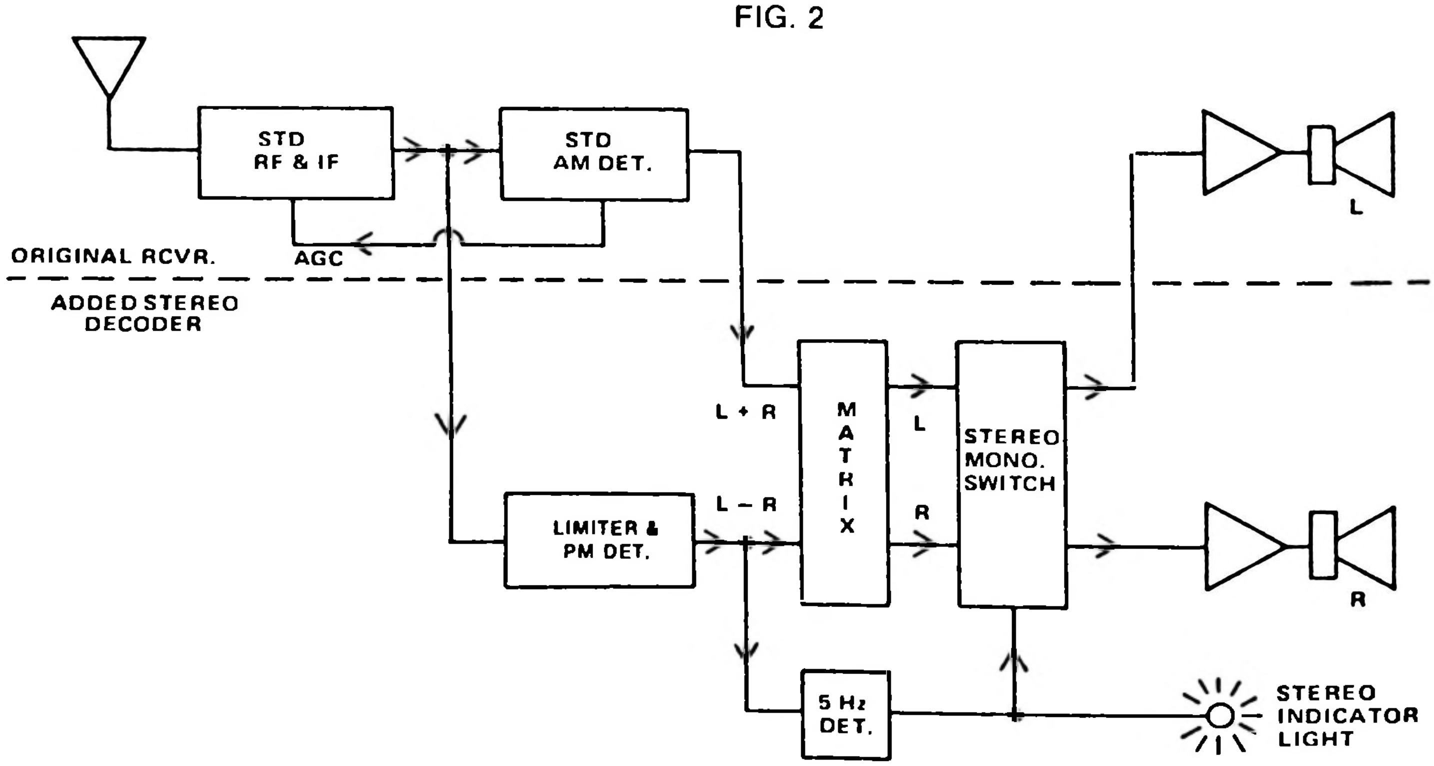


FIG. 2



MOTOROLA C-QUAM AM STEREO SYSTEM

MOTOROLA

CHRIS PAYNE

SCHAUMBURG, ILLINOIS

Many attempts have been made to develop an AM stereo system using quadrature modulation. However, the basic problem with quadrature systems has always been monaural compatibility; that is, high levels of distortion in typical AM receivers. Previous attempts to solve the incompatibility problem have resulted in poor coverage, reduced stereophonic information, poor separation, or a combination of these ill effects. The Motorola C-Quam (Compatible Quadrature) system solves the usual incompatibility problem without incurring losses in range, separation, or distortion.

AUDIO PERFORMANCE

The Motorola system is, in concept, without flaws. This means that the theory tells us that the system is capable of zero distortion, infinite frequency response, and infinite separation. In practice, the performance is limited primarily by the fidelity of the broadcast transmitter and receiver. In the laboratory, frequency response of 20 Hz to 15 kHz, separation of more than 40 dB and distortions of less than 0.3 percent, have been readily attained. In the field, performance has exceeded FM stereo proofs and won the support of major receiver manufacturers.

AUDIO PROCESSING AND PREEMPHASIS

Actual broadcasting experience with the Motorola system has proven that high degrees of audio processing and preemphasis may be used without difficulty. Stations, in fact, report that the received monophonic sound is as good or better when operating with Motorola AM Stereo as with previous monophonic operation.

SPECTRUM

The Motorola AM Stereo system, when operating with typical state-of-the-art audio processing equipment with preemphasis, occupies essentially the same spectrum as does a monophonic station with the same processing. The Motorola exciter has been FCC-type accepted for broadcast use, without any special stereo filter restrictions.

THE C-QUAM TRANSMITTER

The C-Quam transmitter is shown in Figure 1. Note that pure quadrature is generated by taking L + R and L - R and modulating two balanced modulators fed with R.F. signals out of phase by 90 degrees. In this case the 90 degrees phase shift is derived by using a Johnson counter which divides an input frequency (four times station carrier frequency) by four and automatically provides digital signals precisely 90 degrees out of phase for the balanced modulators. The carrier is inserted directly from the Johnson counter. At the output of the summing network, the result is a pure quadrature AM stereo signal. From there it is passed through a limiter which strips the incompatible AM components from the signal. The output of the limiter is amplified and sent to the broadcast transmitter in place of the crystal oscillator.

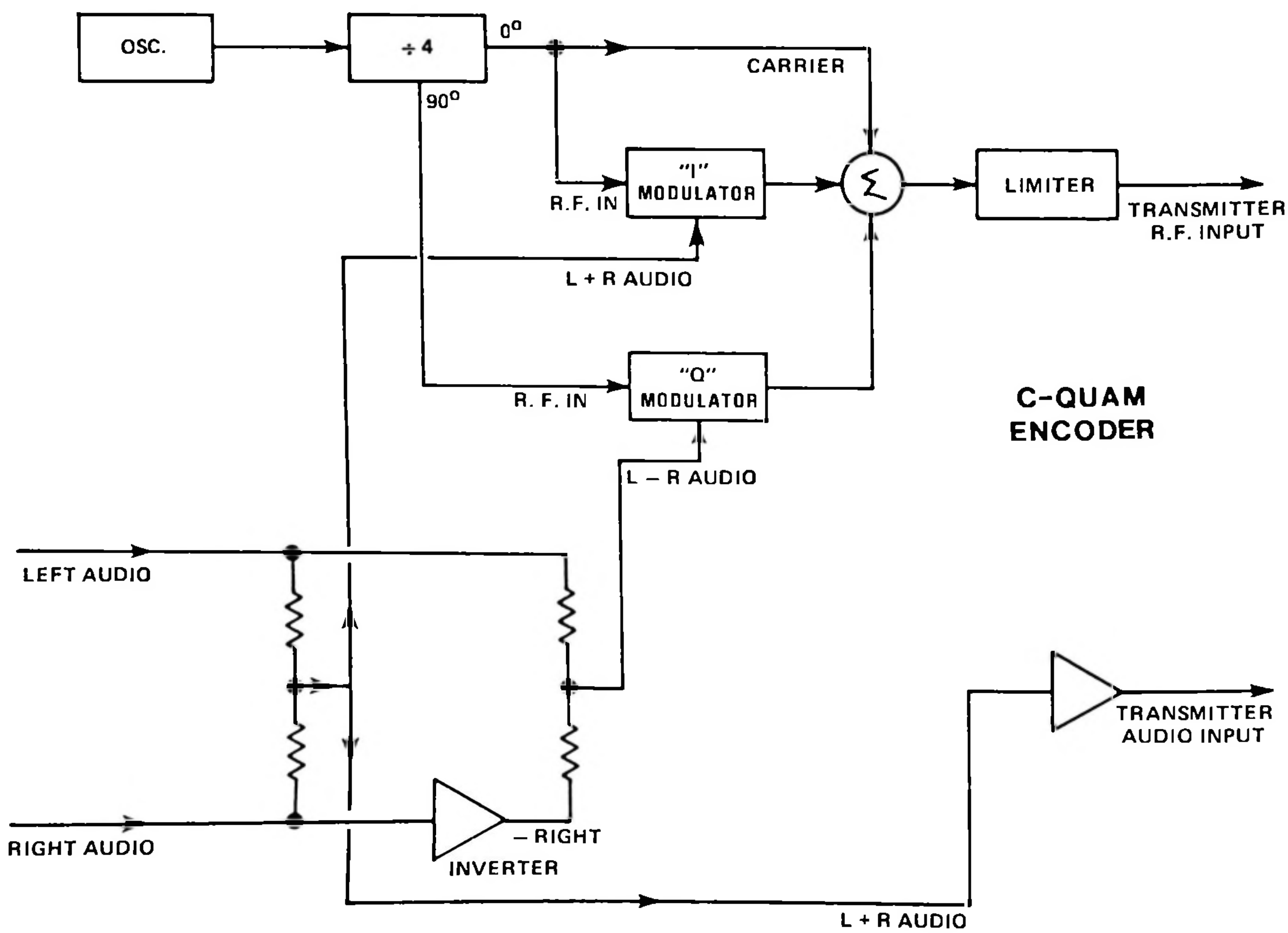


Figure 1.

The left and right audio signals are precisely added and sent as normal and compatible L + R to the audio input terminals of the broadcast transmitter. That's the Motorola C-Quam encoder.

DECODING C-QUAM

C-Quam is decoded by simply converting the broadcast signal (which is already close to a quadrature signal) to pure quadrature and then using a quadrature detector to extract and L - R. Refer to Figure 2. Note that the demodulator contains a section which is a pure quadrature demodulator. In order to prepare the received signal for the quadrature demodulator, it has to be converted from the envelope detector compatible signal that is broadcast to the original quadrature signal that was not envelope detector compatible. This is done by demodulating the broadcast signal two ways; with an envelope detector, and with an "I" detector. The two signals are compared and the resultant error signal is used to gain modulate the input of the "I" and "Q" demodulators.

When the transmitted signal is L + R (monaural, no stereo) the transmitted signal is pure AM or only "I" sidebands. In this case the envelope detector and the "I" demodulator see the same thing. There is no error signal, the input modulator does nothing and the signal passes through without change. However, when a left or right only signal is transmitted, both AM and PM is transmitted and the input signal is shifted in phase to the "I" demodulator and loses some of its "I" amplitude. The envelope detector sees no difference in the AM because of the phase modulation, and when the envelope detector and the "I" demodulator are compared, there is an error signal. The error signal pushes up the input level to the detector. This makes the input signal to the "I" and "Q" demodulators look like a pure quadrature signal and the audio output gives a perfect and L - R signal. The demodulator output is combined with the envelope detector output in a matrix to give left and right audio out.

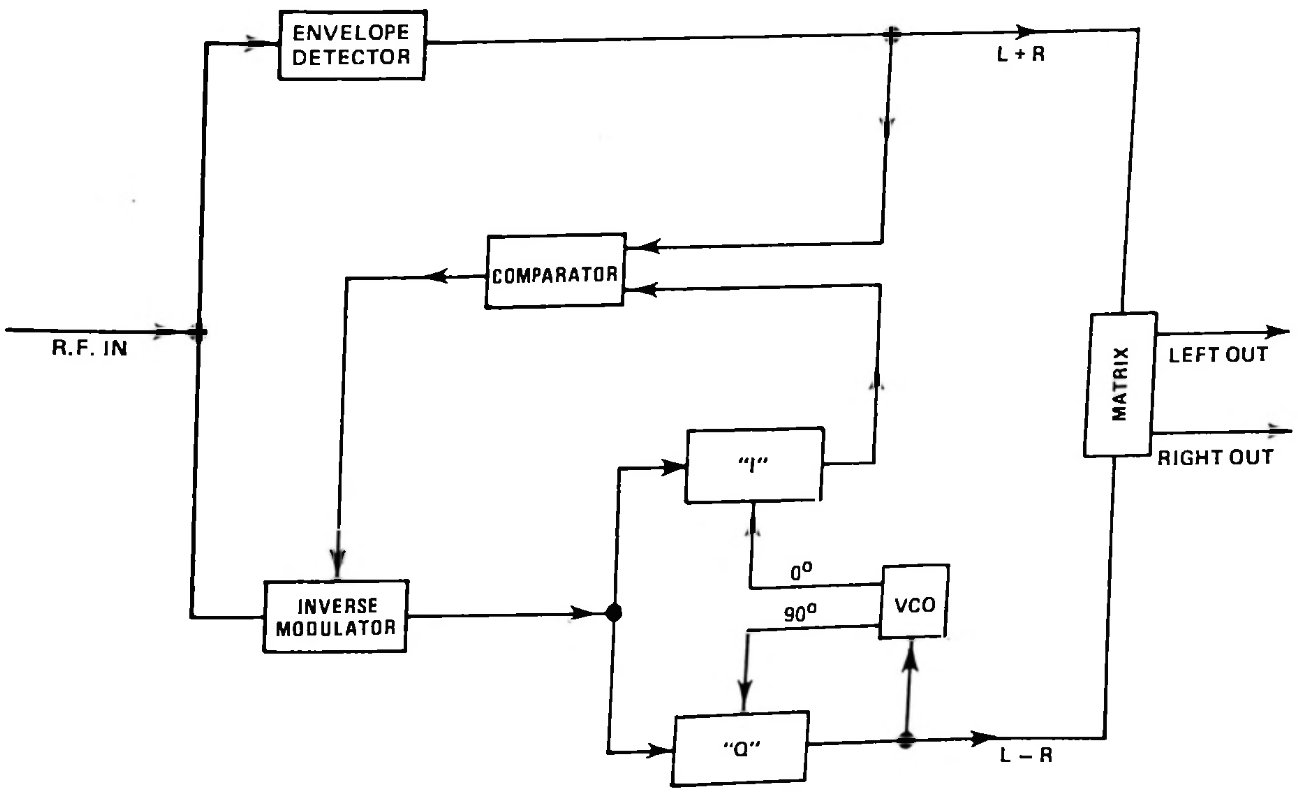


Figure 2.

SYSTEM PERFORMANCE UNDER HEAVY MODULATION

There are many advantages of the Motorola system. One is its performance under 100 percent negative amplitude modulation conditions. When the carrier momentarily goes to zero, as in 100 percent negative modulation, the output of the envelope detector becomes zero. Because of the action of the comparator and inverse modulator, the output of the "I" demodulator goes to zero. Simultaneously, therefore, the output of the "Q" demodulator also is forced to go to zero. This means that there will be no severe noise or interference popping from the stereo channel under 100 percent negative amplitude modulation.

THEORETICAL AUDIO PERFORMANCE

There is no theoretical limit to the audio frequency response, to the lower limit of distortion or to the separation of the Motorola system. All of these characteristics are primarily limited by the performance of the encoder, AM transmitter, and modulation monitor or receiver. The only other limit would be due to the bandwidth of emissions under stereo modulation conditions with single tone modulations above 7500 Hz. It is theoretically possible to force the system with tone modulations, to exceed the FCC bandwidth limits permitted for AM stations by a few dB. However, under the most heavily processed and preemphasized stereo program conditions, the emissions are well below FCC requirements for bandwidth.

ACTUAL STEREO PERFORMANCE

The actual measured performance of the Motorola system connected to a typical plate modulated AM broadcast transmitter is excellent, nearly that of FM stereo. The frequency response of left or right only transmissions is basically the same as the AM response of the transmitter. (In the case of tests made at WIRE in Indianapolis, about 1 dB down from 30 Hz to 12 kHz). The separation was held to better than 30 dB from 100 Hz to 7.5 kHz and within 20 dB from 50 to 10 kHz. Distortion of the left or right only transmissions were below 1.5 percent from about 80 Hz to 7.5 kHz. This data is taken from actual measurements made from an AM stereo modulation monitor.

RECEIVER AUDIO AND COVERAGE

Because the Motorola AM Stereo system acts like a full quadrature system under normal stereo conditions, its signal to noise characteristics are excellent. Under typical stereo programming, the increase in noise is estimated to be about 1.5 dB which is nearly imperceptible. In other words, the stereo coverage of the Motorola system is essentially the same as the monaural AM coverage.

The audio fidelity that can be expected from AM stereo radios depends upon the receiver design. In the past, receiver manufacturers were generally not willing to make AM radios with better fidelity because they felt that there was no demand for such equipment. With AM stereo, the ability to offer "stereo" means a much more attractive product which will allow a high price. Therefore, the AM stereo radio designer will have a much wider latitude in designing the receiver selectivity and will be more able to achieve higher fidelity while holding down the susceptibility to adjacent channel interference. With slightly more expensive I.F. selectivity, the receiver designer can obtain audio response well beyond 5 kHz while still rejecting the adjacent channel interference. Two bandwidth receivers, variable bandwidth receivers, and fixed bandwidth receivers to 7 kHz may be excepted for AM stereo.

CONCLUSION

The Motorola AM Stereo system is the system designed to meet the needs of AM radio. In other words, designed to perform well yet to be received inexpensively so that millions of radio listeners can enjoy AM stereo. While the design is adaptable for cost effective radios, it is also the best overall technical performer for the broadcaster because it is capable of FM-like audio performance with full amplitude modulation capability and stereo coverage essentially the same as monaural coverage.

BELAR AM STEREO SYSTEM

ARNO M. MEYER

BELAR ELECTRONICS LABORATORY, INC.

DEVON, PENNSYLVANIA

The Belar AM-FM System is a matrixed system which uses left and right channel audio material to form a sum signal (L+R) and a difference signal (L-R). The sum signal is applied to the amplitude modulation circuitry of a conventional AM transmitter, as currently done in monophonic broadcasting. Stereo (L-R) information is used to angularly modulate the AM carrier. The angular modulation has FM characteristics for low audio frequencies, and PM (phase modulation) characteristics for mid audio frequencies. The L-R signal from the audio matrix circuit is applied to a pre-emphasis circuit. This FM-to-PM changeover is accomplished with a controlled pre-emphasis network. Pre-emphasis, combined with de-emphasis in the receiver, serves to reduce the detected noise in the stereo signal; primarily for the low audio frequency range. The Belar system has a significant advantage over competing systems in this regard. Peak FM deviation of the lowest audio frequencies transmitted is 312.5 Hz or 6.25 radians at 50 Hz. Similarly, the phase modulation index of the highest audio frequencies is 0.85 radians/sec. The pre-emphasis network has a controlled time constant to limit deviation at high audio frequencies, which will limit the occupied bandwidth of the stereo signal. A 10 Hz tone with a deviation of 2 radians is used as an indicator for receivers of the transmission of a stereo signal.

The receiver is essentially a complementary system to the transmitter. A conventional diode detector detects the L+R signal in the envelope. A limited IF signal (free of AM modulation) is applied to an FM discriminator. An audio high-pass filter rejects the pilot tone. The de-emphasis network is the complement of the pre-emphasis network in the transmitter. The L+R and L-R signals are applied to an audio matrix which produces the original left and right audio signals. The output of the discriminator is applied to a bandpass filter centered at 10 Hz so that most program material and noise is removed from around the pilot tone. The presence of a signal at the output of this filter triggers a stereo indication mechanism.

The use of transmitted pre-emphasis and a de-emphasis receiver (as is done in the Belar system) will reduce the low frequency microphonic and tuning noise which occurs naturally in non-synthesized (and some synthesized) receivers. This is one of the significant advantages of the Belar systems, and is one of the reasons the Belar system received a #2 rating in the FCC original analysis (FCC "Further Notice").

How International Spectrum Decisions

Affect Broadcasting in the United States

A. James Ebel

Cornhusker Television Corporation

Lincoln, Nebraska

How International Spectrum Decisions Affect Broadcasting in the United States

International spectrum decisions are made at either World Administrative Radio Conferences (WARCs); or at Regional Administrative Radio Conferences (RARCs); or through multi-lateral agreements between two or more countries with common interests in radio communication. WARCs, RARCs, and some multi-lateral agreements are negotiated under the auspices of the International Telecommunications Union (ITU) -- the oldest international organization still in existence.

The International Telecommunications Union

Just a word or two on the history of the ITU. Signalling at a distance has been the goal of mankind over the ages. Early circuits were started with a visual telegraph -- a series of semaphore towers which could encode messages through the various positions of the arms. The encoded messages could be seen at a distant semaphore tower by means of telescope and could be re-transmitted on down the line by that semaphore tower. A network of such optical telegraph circuits covered most of France early in the 19th Century.

The electro-magnetic telegraph was invented almost simultaneously in England by Cooke and Wheatstone and in the United States by Samuel F. B. Morse. The development of telegraphy circuits with repeaters made possible rapid transmission of messages over relatively great distances. Because the telegraph was initially used by the railroad and the fact that railroad trains frequently crossed national borders it was essential to be able to communicate internationally. Initially, this was done by setting up a translation point at the border. Messages received from within a country, could be re-transmitted into the adjacent country using its language and the special telegraph code they had developed. Not only were languages different but the telegraph codes were different. It became obvious over the years that some international agreement was necessary for cross-border transmissions. So in 1865 in Paris the International Telegraphic Union was formed. They adopted the Morse code as a standard and set up international charges for message traffic. With the development of radio transmission by Hertz, Marconi, DeForest and others, the need for an International Radio Telegraph Union also became obvious. It was necessary to establish frequencies on which signalling would take place, particularly because ships that travelled world-wide would have to be able to communicate with ports any place in the world. International distress frequencies had to be established and interference had to be considered. So, a Radio Telegraph Union was established in 1903. The development of sound broadcasting increased the importance of the Union because radio broadcast signals did not recognize country borders and interference became rampant. The International Telegraphic Union's Conference and the Radio Telegraph Union's Conference in 1932 were held together in Madrid, Spain. Out of this meeting came a merger into a single organization -- the ITU. Subsequently, the name was changed to the International Telecommunication Union to reflect the fact that the organization was involved in all aspects of electrical communications systems; telegraph, telephone, television, radio and satellite.

Functions of the ITU

The ITU Convention governs the operation of the organization. This Convention is an international treaty agreement which sets up the structure, and operation of the ITU which has its headquarters in Geneva, Switzerland. All nations which are signatories to the Convention are bound under International law to observe all ITU regulations. The Convention is updated at regular Plenipotentiary Conferences. A 36-member Administrataive Council is elected by the Plenipotentiary Conference and is responsible for guiding the work of the ITU between Conferences.

The ITU under the Administrative Council consists of four working bodies as follows:

1) The General Secretariat -- which handles executive management of the Union;

2) The International Frequency Registration Board (IFRB) which supervises the recording of frequencies and orbital positions and gives advice in spectrum management;

3) The International Radio Consultative Committee (CCIR) which studies the technical basis for the various systems of radio communications; and prepares reports and recommendations; and publishes "green" books at regular intervals which set forth international agreements on technical subjects. Material from the "green" books is used as a basis for negotiation at World Conferences.

4) The International Telegraph and Telephone Consultative Committee (CCITT) which studies technical, operational and tariff questions relating to telephony and telegraphy.

The ITU is financed by a class-unit system first adopted by the Vienna Conference in 1868. Each member freely chooses its class, i.e. the percentage of the total budget expenditures it will contribute in percentage units. Today the classes range from 30 units down to one-half unit. As might be expected the United States is the largest contributor to the support of the ITU.

At each World or Regional Administrative Conference each nation has only one vote no matter how large a percentage of the expenses of the ITU they contribute. Because of the proliferation of developing nations and the fact that developing nations tend to vote as a block, this "one country one vote" policy creates substantial difficulties for the developed nations. At the 1983 RARC the United States vote will have no greater importance than the votes of the smallest countries in the western hemisphere. The success we achieve at International Conferences is due largely to the overwhelming power of our technology and to the technical ability of the members of our delegation.

ITU Technical Studies

Of greatest interest to the technical community is the work of the CCIR. The CCIR study groups are staffed by skilled technical volunteers who meet on a regular basis to study problems referred to it by World Administrative Conferences and to study other technology which appears to need consideration. Study groups are at work in many of the ITU member countries. Papers developed as the result of the work of individual study groups are presented to Interim Working Parties (IWP) and to CCIR Prepartory meetings for consideration. In their final form they wind up as part of the "green" books. The study groups of greatest interest to broadcasters are Study Group 5, Propagation Non-ionized Media; Study Group, 10 Broadcasting Service (sound); and Study Group 11, Broadcast Service (TV). Sub-groups of study groups 10 and 11 referring to satellite broadcasting are of particular interest in view of the upcoming Regional Administrative Radio Conference (RARC). Materials developed by this group will be used as a basis for RARC negotiations, e.g. Antenna paterns; Propagation losses; Interference calculations; and signal-to-noise calculations. These are a few among the many subjects agreed upon by the CCIR.

Recent Spectrum Allocation Changes

The 1979 General World Administrative Radio Conference (WARC) made a number of decisions that will affect the future of Broadcasting in the United States. In Low-frequency broadcasting something that didn't happen will affect us. There was a proposal for adding low-frequency broadcasting to the Table of Allocations for Region 2 (North and South America are in Region 2). This proposal did not make its way into the United States Position and was not successfully proposed by any other Region 2 country so it didn't happen. The proposal was made by Educational interests and opposed by those who use power-line communication.

In the standard broadcasting band -- which is the MF broadcasting band -- the lower end was extended by one channel, which is to be local channel. The upper end of the band was extended 10 channels subject to a plan to be set up by a Regional Administrative Radio Conference in 1985. The first two channels would be available by July 1, 1987; and the remaining 8 channels would become available July 1, 1990. This delay is to allow time for orderly transition.

There were no major changes affecting TV broadcasting in the United States in the VHF band. There was a minor change in the space between Channel 4 and Channel 5 to clean up the problem of some unallocated spectrum. There was also no change in the FM band.

In Region 2, in which the United States is a part, the band limits of the UHF band remain the same -- although internally the FCC has decided to allocate Channels 70 through 83 to Land Mobile. The United States also wanted to affect sharing the UHF band by Land Mobile on a primary basis. This was unacceptable to other nations at the 1979 WARC. The final wording called for Land Mobile sharing subject to agreement obtained under procedures set forth in Article N-13A which provides for a supplementary procedure to be applied in cases where a footnote in the Table of Frequency Allocations requires an agreement with an Administration. The United States considered this procedure objectionable because it was too cumbersome and time consuming to allow the establishment of Fixed Service and Mobile Service operations in this band on a shared basis with UHF-TV broadcasting.

In the 12 GHz band in Region 2 Terrestrial broadcasting was allocated in the 12.2-12.7 GHz band sharing with other services including broadcasting satellite on a primary basis. However, the international allocation did not affect the FCC allocation. The FCC has removed terrestrial broadcasting entirely from the 12 GHZ band.

How the ITU Affects US Broadcasting

This brings us to the bottom line of what this paper is all about...."How International Spectrum Decisions Affect Broadcasting in the United States". Spectrum decisions in the United States will be affected only when the spectrum decision involves the possibility of international interference. If the establishment of a service will not create interference outside of the borders of the United States the service can be established irrespective of what is in the International Table. For example Land Mobile Sharing in the UHF band can take place anywhere except inside of the 250-mile distance to the border. Another example would be the listing of Terrestrial broadcasting in the Table of Allocations for the 12 GHz band in Region 2. This does not mean however that the FCC has to make such an allocation in the United States.

Starting June 13 in Geneva Switzerland the 1983 RARC Conference for Region 2 will get underway. This Conference is to set up a plan for broadcasting by satellite in Region 2. It is difficult to say how we will make out in these negotiations. Whatever happens will have an effect on broadcasting in the United States. A very direct effect on satellite-to-home broadcasting; and an indirect effect on terrestrial broadcasting, through the introduction of a new competing delivery system.

national association
of broadcasters