Digital Audio for NTSC Television

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0. Abstract

A previously proposed method of adding a digital carrier to the NTSC broadcast channel was found to be marginally compatible with adjacent channel operation. The technique also has some problems unique to broadcasters. Digital transmission techniques are reviewed, and a new set of digital transmission parameters are developed which are thought to be optimum for digital sound with NTSC television.

I. Introduction

It is very feasible to compatibly add digital audio to the NTSC television signal as carried on cable television systems. Besides the marketing advantage which can accompany the use of anything "digital", there are some real advantages to digital broadcasting, especially where the transmission path is imperfect. Digital transmission is inherently robust. While the coding of high quality audio into digital form theoretically entails a loss of quality, there is no further loss of quality when the digits are transmitted through an error-free channel. This is in stark contrast to the transmission of analog audio, in which a "perfect" channel is required to avoid degradation. Unfortunately, the intercarrier sound channel used with broadcast television is somewhat less than perfect, and inherently limits the sound quality of the analog BTSC stereo system.

Digital transmission techniques have matured to the point where they can be economically applied to broadcast audio. Previous work[1,2] led to a proposal for the addition of digital audio to the NTSC television broadcast signal. That proposal borrowed heavily from work performed in Sweden[3] and Finland where a 512 kb/s QPSK carrier was extensively tested with PAL system B. The similarities between B-PAL and M-NTSC indicated that the Scandinavian test results would apply in the U.S. Our original 1987 proposal was:

A. QPSK carrier with alpha=0.7 filtering.

B. Carrier frequency 4.85 MHz above video carrier.

C. Carrier level -20 dB with respect to peak vision carrier level.

Compatibility testing of that system has been performed, and some television sets have been found on which the data carrier causes detectable interference to the upper adjacent video channel in a clean laboratory setting. It should be noted that with these problem sets, the FM aural carrier also caused noticeable interference to the upper adjacent picture. The interference from data occurs into luminance, and manifests itself as additive noise between approximately 1 MHz and 1.4 MHz. The noise level in a problem TV is subjectively similar to the noise level resulting from a video carrier-to-noise ratio (CNR) of approximately 47 dB. While this level of interference may not be detectable given the current quality level of cable signals in consumers' homes, the potential of optical fibre to raise the quality level of cable TV warrants a reduction in the level of interference by approximately 6 dB to be safe.

The system should be compatible with broadcast television as well as cable television. We must consider broadcast problems which can result from existing transmitter plant, and broadcast spectrum allocations. These considerations also call for some changes to the previously proposed system.

II. Digital Modulation Basics

Figure 1 shows the spectrum of a random data stream with a data rate of 250k bits/sec. In a digital transmission the data is always intentionally scrambled so that the transmitted data appears random. Note the spectral nulls at multiples of the data rate (250 kHz, 500 kHz, etc.). From Nyquist sampling theory we know that all information in this signal may be gleaned from the first 125 kHz, i.e.

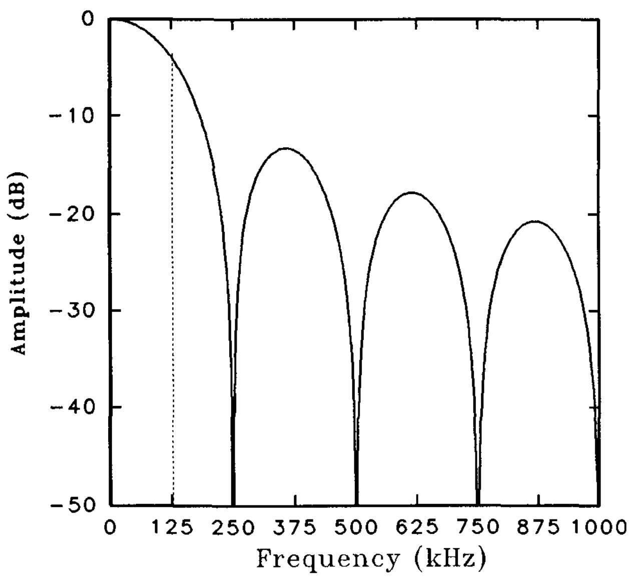

Figure 1 Spectrum of 250 kb/s data

the Nyquist frequency which is ½ of the 250 kHz sample rate. If we brick wall filter this signal at the Nyquist frequency of 125 kHz and use it to modulate a carrier, we will have a double sideband modulated bandwidth of 250 kHz, which will carry 250k bits/sec of data (1 bit/sec/Hz). If we add a second similar channel at the same frequency but in quadrature, we will double the information in the 250 kHz bandwidth RF signal to 500k bits/sec, or 2

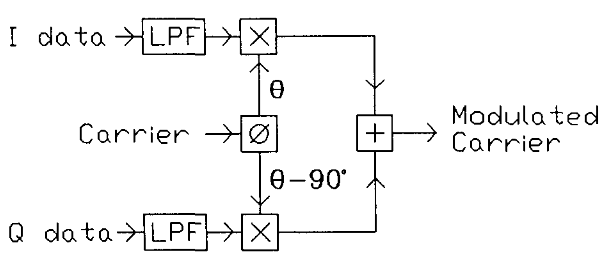

Figure 2 QPSK Modulator

bits/sec/Hz. Fig. 2 shows a block diagram of this modulator which is simply a pair of double sideband AM modulators in quadrature. The double sideband modulator can be considered to be a phase modulator, generating phases of 0 and 180 degrees. The second modulator can be considered to generate phases of 90 and -90 degrees. The combination of the two phases will generate one of four phases, ± 45 and ± 135 degrees. This form of modulation is typically called **Q**uadrature **P**hase **S**hift **K**eying or **QPSK**.

Fig. 3 shows the constellation diagram for QPSK. There are four possible states of the RF carrier phase. We refer to a chosen state as a symbol. Since there are four possible states per symbol we are transmitting two bits per symbol with a symbol rate of 250k symbols/sec, for a total data rate of

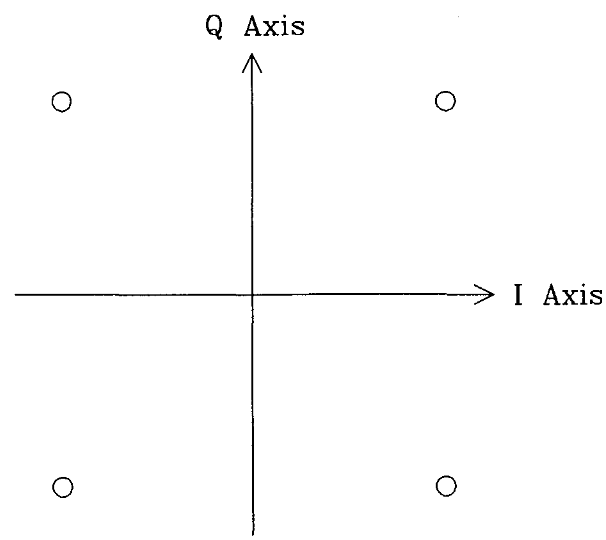

Figure 3 QPSK Constellation

500k bits/sec. If the low pass filters were not present, the phase would be instantaneously jumping between the four states shown, at a rate of 250 kHz. In practice the actual signal will be continuously traversing between these points and the receiver will sample the signal at the time when the signal passes through the constellation points.

The QPSK receiver is shown in Fig. 4. A carrier recovery loop coherently regenerates the transmitted carrier, which was suppressed in the modulator. Quadrature mixers demodulate the I and Q signals, low pass filters limit the effective receive bandwidth,

$$H(j\omega) = \begin{cases} 1, & \left[0 \le \omega \le \frac{\pi}{T_s}(1-\alpha)\right] \\ \cos^2\left\{\frac{T_s}{4\alpha}\left[\omega - \frac{\pi(1-\alpha)}{T_s}\right]\right\}, & \left[\frac{\pi}{T_s}(1-\alpha) \le \omega \le \frac{\pi}{T_s}(1+\alpha)\right] \\ 0, & \left[\omega \ge \frac{\pi}{T_s}(1+\alpha)\right] \end{cases} \qquad (1)$$

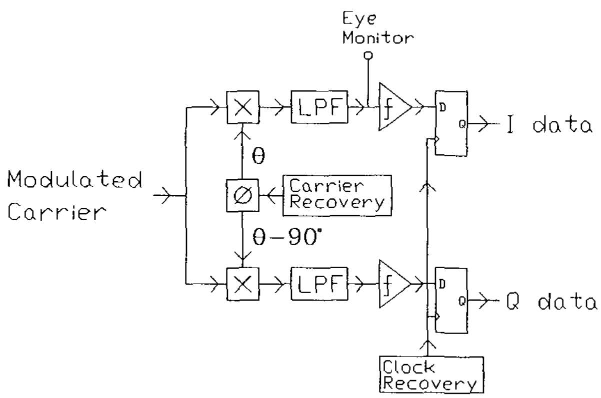

Figure 4 QPSK Demodulator

and comparators and clocked latches recover the data. The latches are driven by a symbol timing recovery circuit which reconstructs the transmit data clock at the proper phase so that the sampling time is correct.

While QPSK can theoretically achieve a spectral efficiency of 2 bits/sec/Hz, this is not achievable in practice since a brick wall filter is not realizable. The steepness of the filter determines the spectral efficiency. Steep filters ring, and as the spectral efficiency of QPSK is raised by sharpening the filters, the ringing of the individual data pulses increases. This ringing can cause the pulses to interfere with each other (intersymbol interference) unless the ringing is controlled. If the filtering is done properly, each individual pulse waveform, except for the pulse being detected, will pass through zero at the instant the receiver samples the waveform. Nyquist filters provide the desired characteristics. A commonly used class of Nyquist filters are the "raised cosine" filters which have a frequency response as described in Eq. 1, and shown in Fig. 5. The alpha term determines the fractional excess bandwidth over a perfect brickwall (alpha=0) filter. A filter meeting the amplitude response of Eq. 1, and having constant group delay (linear phase) will have no intersymbol interference.

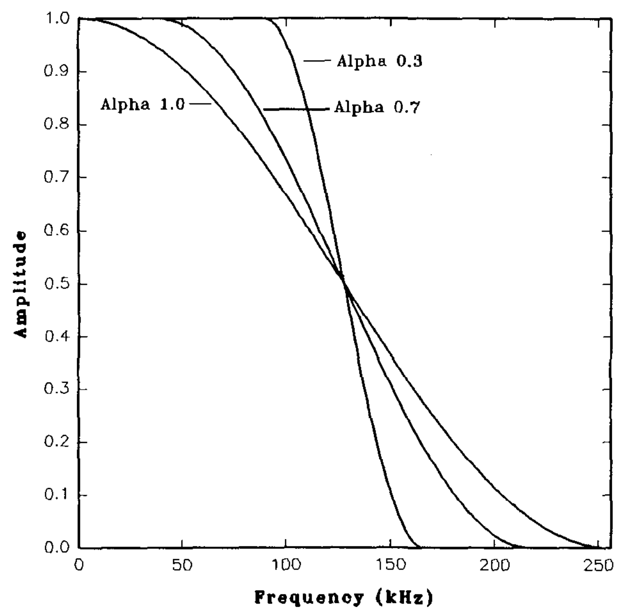

Figure 5 Raised Cosine Nyquist Filters

When we speak of the channel filter, we mean the filter function of the complete channel. This includes the transmit baseband lowpass filter, any transmit RF filtering, receiver preselection and IF filtering, and receive baseband lowpass filtering. It also includes the effect of the transmission path which might have a non-flat response (or selective fade). Typically, all of the intended filtering is designed into the system at IF or baseband, and all other circuitry is made wideband. The most precise control over filtering is achieved with baseband lowpass filtering. It is much easier to make a precision 125 kHz lowpass filter than a precise 250 kHz IF filter.

$$QPSK\ Spectral\ Efficiency = \frac{2}{1+\alpha}\ bits/Hz \qquad (2)$$

As alpha is reduced, spectral efficiency of QPSK is improved. Equation 2 shows the relationship between spectral efficiency and alpha. As alpha is reduced, the data waveform will have more ringing and the "eye" pattern will become more complex. The eye pattern is observed at the point where the analog waveform is sampled to recover the data. In the QPSK decoder block diagram (Fig. 4), the eye monitor point is shown for the I channel, and is just after the baseband lowpass filter. In Fig. 6 we see some examples of eye patterns for alpha=0.3, alpha=0.7, and alpha=1.0.

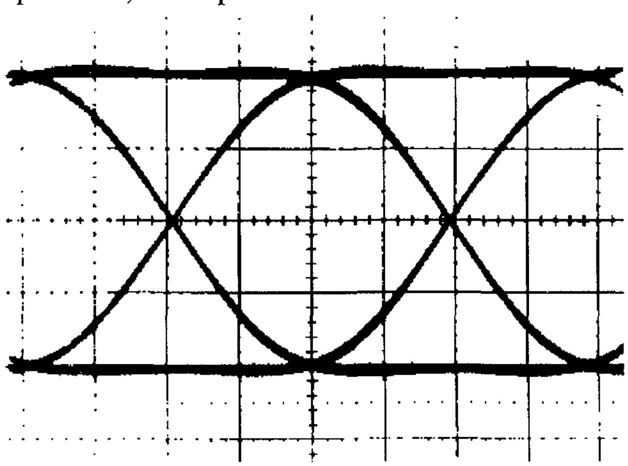

Figure 6a Eye Pattern, Alpha = 1.0

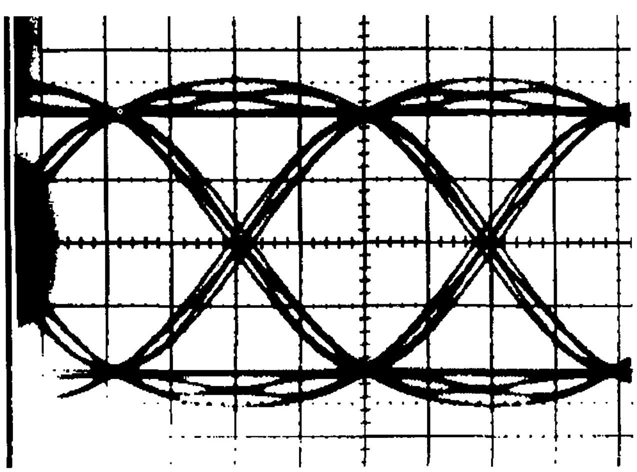

Figure 6b Eye Pattern, Alpha = 0.7

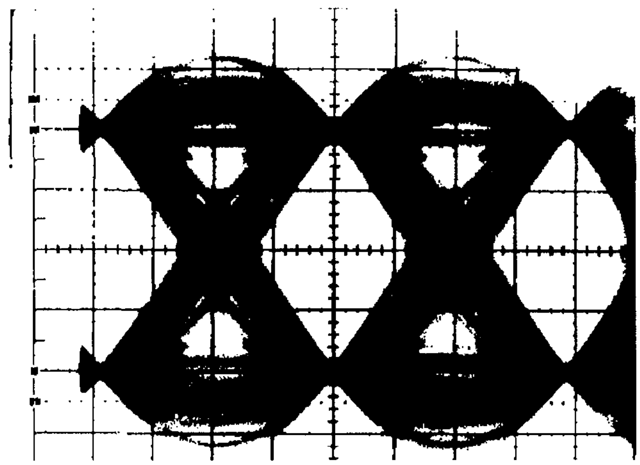

Figure 6c Eye Pattern, Alpha = 0.3

The eye patterns become more complex as alpha is lowered and spectral efficiency is improved. More accuracy is required in implementation of low alpha systems. Any error in amplitude or phase linearity will rapidly degrade a complex eye. The eye will appear to close. Fig. 5d shows an alpha = 0.3 eye with a lot of closure due to poor filtering. Partial eye closure due to imperfect filtering leaves less margin against noise and interference.

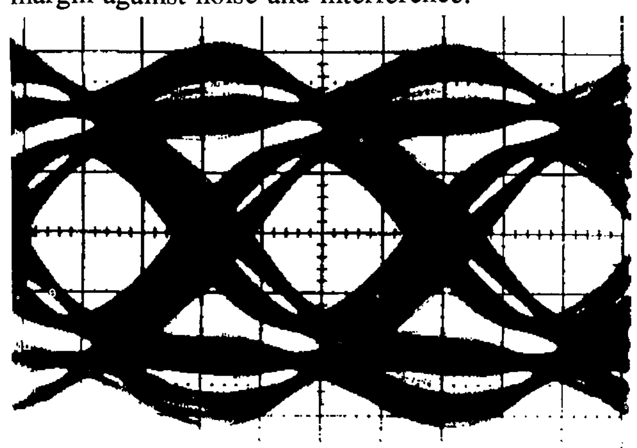

Figure 6d Alpha = 0.3, Poor Filter Accuracy

For optimal performance over a noisy channel, the amplitude portion of the filtering should be partitioned in equal portions between the transmitter and receiver i.e. the transmit filter and receive filters should both have an amplitude response equal to the square root of the nyquist filter response. The cascade of the two filter characteristics will then be a Nyquist filter. The combination of the filters should have linear phase (constant group delay). If both the transmit and receive filters individually have linear phase, the combination will as well. However, it is also acceptable for the receive filter to have group delay variations and to compensate for these in the transmit filter.

Attempting to achieve good spectral efficiency with low alpha filtering requires the use of more accurate filters with good phase compensation. Filters may be implemented with either analog or digital circuitry. Since we are starting with digital data, the use of a digital filter in the transmit side is attractive. The receiver is more easily implemented with an analog filter which does not require the use of A-D converters with high sample rates and anti-aliasing pre-filters. In a broadcast application we want to keep the receiver cost low, so we want to use the minimal receive filter (without phase compensation). If the receive filter magnitude response accurately matches the square root of a Nyquist response, then the transmit filter can be implemented as an FIR (finite impulse response) filter with an impulse response which is simply the time reversal of the impulse response of the receive filter. The cascade of the filters will then have a symmetrical impulse response with constant group delay, and will have an amplitude response equal to that of a Nyquist filter. The matched filter criterion is satisfied, so performance in the presence of noise will be optimal. All phase compensation is done in the transmit filter so receiver cost is minimized.

In order to achieve spectral efficiencies of 2 bits/sec/Hz or greater, QPSK must be abandoned and a higher level modulation method used. This can be done by hitting the modulator filters (Fig. 2) with multilevel data symbols. With QPSK, the I and Q data are simple binary symbols with two levels (1 and 0, or +1 volt and -1 volt). If pairs of bits are fed to 2 bit D-A converters, we can form four level data (0, 1, 2, 3; or +3 volts, +1 volts, -1 volts, -3 volts). QPSK extended to 4 level symbols is known as 16 QAM (Quadrature Amplitude Modulation). 16 QAM has a theoretical (with ideal brick wall filtering) spectral efficiency of 4 bits/sec/Hz. Unfortunately, the added spectral efficiency is achieved at the expense of ruggedness. The eye pattern of 16 QAM is similar to that of three QPSK eyes stacked on top of one another. For the same absolute eye opening (which would give the same performance in a noise
or interference environment), the peak level is 3 times higher, or about 9.5 dB. For the same carrier level, the 16 QAM signal is 9.5 dB less rugged than QPSK.

III. Partial Response Signaling

Another attractive technique which can achieve improved spectral efficiency is known as "partial response signaling". This method involves the use of a controlled amount of intersymbol interference that is removed by an extra processing step in the receiver. The simplest form of partial response involves performing a running average of adjacent pairs of bits before driving the modulator. If we average a pair of bits which can each be +1 or -1 we get symbols with three levels: +2, 0 or -2. This 3-level signal still only contains 1 bit of information per baseband symbol, but the spectral characteristics are changed. The spectrum of the data is changed

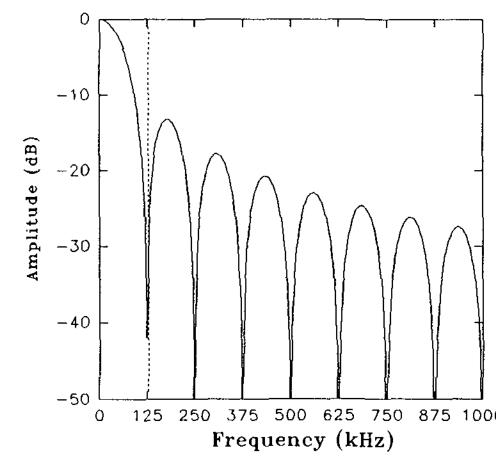

Figure 7 Partial Response Spectrum

from that shown in Fig. 1 to that shown in Fig. 7. The first spectral null has moved from 250 kHz down to 125 kHz. If we filter off all of the energy beyond 125 kHz (which is relatively easy), we will achieve a spectral efficiency of 1 bit/sec/Hz at baseband. The use of a pair of these partial response system modulators in quadrature is known as QPRS. The constellation diagram for QPRS is shown in Figure 8.

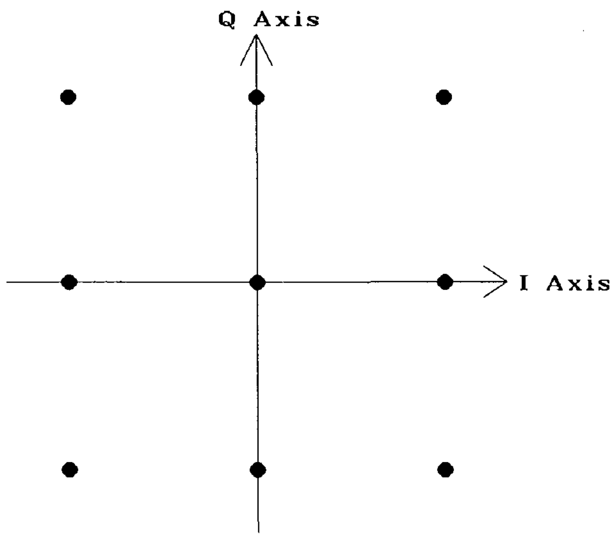

Figure 8 QPRS Constellation

The pairwise bit averaging is equivalent to a filter with an amplitude response equal to a cosine function, as shown in Fig. 9. In order to have matched filtering, the bit averaging should not be done at the transmitter, but the cosine filter function should be divided equally between the transmit and receive filters. That is, both the transmit filter and the receive filter should have an amplitude response equal to the square root of the first quadrant of a cosine. The transmit filter may be a digital FIR filter with an impulse response which is the time reversal of that of the receive filter.

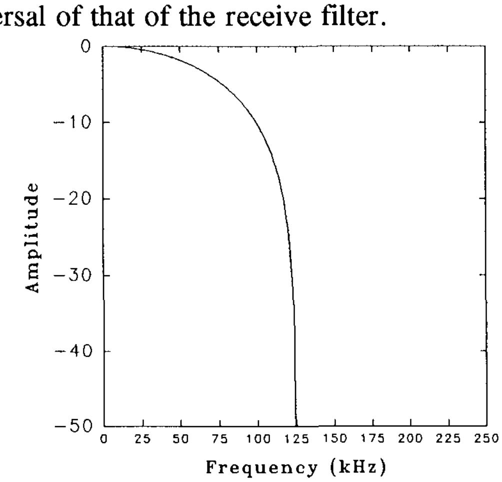

Figure 9 PRS Cosine Filter

Fig. 10 shows the spectrum of the transmitted data carrier for several modulation schemes. The narrowest spectrum is for QPRS, the second narrowest is alpha=0.3 QPSK, and then we have

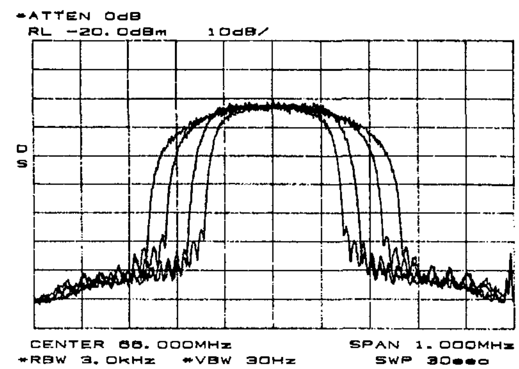

Figure 10 Comparative RF Spectra

alpha=0.7 and alpha=1.0. The transmit filters used are FIR filters with a magnitude response equal to the square root of the raised cosine curve (for the QPSK cases), or the square root of a cosine curve (the QPRS case). The spectrum of QPRS is very attractive, but its eye pattern (Fig. 11) shows it to be less rugged than QPSK. Since there are two eyes, in order to achieve the same absolute eye opening (and thus the same ruggedness), the peak power must be increased by a factor of two, or 6 dB. For the same peak power, QPRS should be 6 dB less rugged. There are some mitigating factors though.

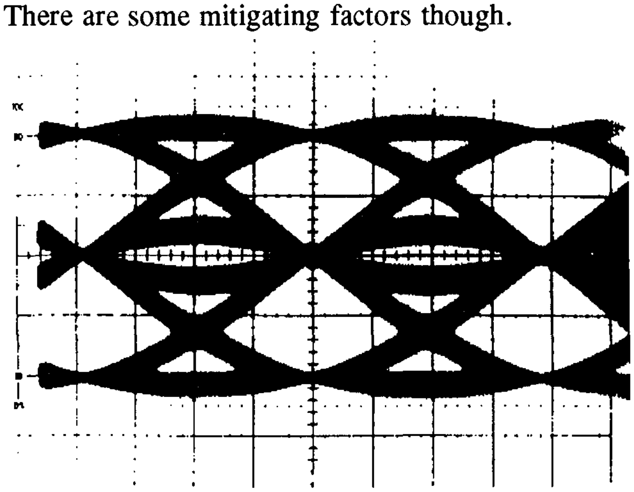

Figure 11 PRS Eye Pattern

First, the PRS data stream spends half its time at the 0 level (data = 1,0 or 0,1), and half at the +2 or -2 level (data = 0,0 or 1,1). This means that half of the time the carrier is modulated by zero, and half of the time it is modulated by the peak value. QPRS will have an RMS power level half that of QPSK for the same peak level. For the same RMS power level, QPRS will thus lose only 3 dB of ruggedness compared to QPSK, and not the 6 dB that is lost with equal peak power levels. Second, since there is some redundancy in the PRS 3-level data stream, it is possible to detect and correct some errors without intentionally including redundant FEC data in the data multiplex. Coding gains of 1.5 to 2.0 dB can be achieved economically. The combination of these factors indicate that for comparable carrier-to-noise ratios (CNR), QPRS is only about 1 to 1.5 dB less rugged than QPSK against gaussian noise. QPRS is an attractive candidate if its narrow spectrum would solve our interference problem.

While QPRS requires some added complexity in the receiver, the amount is within reason for a consumer product. If the narrowness of the QPRS signal significantly reduces the potential for interference, QPRS could be adopted for digital TV sound broadcasting without seriously impacting the receiver cost.

IV. Practical Broadcast Considerations

In the case of adjacent channel operation, the proposed data carrier is placed above the FM sound carrier of one channel, and below the lower vestigial sideband of another channel. Figure 12 shows the RF spectrum of TV Ch 3 and 4, with an Alpha = 0.3 QPSK carrier inserted between the channels. In the case of adjacent channel operation, it makes little difference whether we say that the data signal is assigned to the upper or the lower channel. The BBC[4] choose to allocate the signal to the lower channel. There are advantages to this choice.

Depending on the actual vestigial sideband filtering in the video modulator, there may be some overlap of video energy into the spectrum occupied by the data signal. This will allow video to interfere with the data, possibly causing data errors. In the case of broadcasting without adjacent channel operation, placing the data carrier at the top of the channel means that ones own video will not interfere with ones own data.

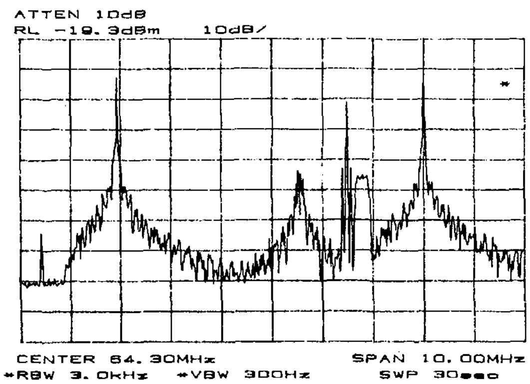

Figure 12 Ch3, Data, Ch4 Spectrum

There is also an advantage in receiver complexity. If the first sound IF is widened slightly, the inter-carrier mixer used to recover the 4.5 MHz FM sound will also recover the data sound carrier at 4.85 MHz. This point can be tapped and fed to the QPSK demodulator. If the data carrier is associated with the upper channel, the inter-carrier offset will be -1.15 MHz, and additional IF filtering and another inter-carrier mixer will be required. Filtering will have to be sufficient to reject the image at 1.15 MHz above vision. The choice of placing the carrier above the channel is a good one, and we originally agreed with it.

Unfortunately, there are some serious problems with this choice for broadcasters in the U.S. The most serious is that in the case of TV Ch. 6, the data carrier would lie at 88.1 MHZ, making both 88.1 and 88.3 unusable for FM broadcasting. The second is that many transmitter installations are unable to easily broadcast the signal due to the notch diplexers in use. Stations which use separate amplification for sound and vision would introduce the signal into the sound transmitter. The sound and vision transmitters are combined in a device known as a "notch diplexer" in which tuned cavities (creating notches) reflect the sound signal into the antenna. There are no cavities tuned to the frequency of the data signal, and the diplexer would need expensive modifications in order to add them. While this is not a technical problem, it is an economic one and could seriously hamper the acceptance of the digital system by broadcasters.

Another problem, unique to cable, has to do with sync suppression scrambling. When sync information is carried on the FM sound carrier as pulse modulation, the spectrum of the FM sound carrier is widened and overlaps the data spectrum. The presence of the data may cause misbehavior in the descrambler, and the presence of the sync information on the FM sound carrier may cause errors in the demodulated data.

Choosing to place the data carrier below the vision carrier appears to be the correct choice for the U.S. Broadcasters may add the signal to the vision transmitter just after the vestigial sideband filter, and it will be amplified linearly along with the picture. At this frequency there will be no problem with the notch diplexer. With the carrier below the channel, there is no problem with a scrambled channel, although the same problem will exist if the FM sound carrier on the lower adjacent channel is pulse modulated.

V. Interference of Data into Picture

Placing the data signal on the lower sideband of our own channel makes it even more critical to avoid interference into video because now we will interfere with ourselves. The mechanism of interference is imperfect VSB Nyquist filtering in the TV receiver. While it is easy to make a TV receiver reject the data signal there are a wide variety of receivers in use, some of which have poor rejection. Fig. 13 shows the ideal receiver IF filter response. Also shown is the kind of response that could be susceptible to interference. The problem is the 'tail' in the rolloff. If the filter goes right down to near zero response at 1 MHz away from the vision carrier the data signal will be fully rejected. If the filter has a tail, the data will not be rejected and may be visible in the picture.

The only way to reduce the potential interference is to: 1) reduce the data carrier level, and 2) move the data carrier farther away from the video carrier (so as to get farther down on the slope of the receive filter). The data carrier can be moved by either moving the center frequency or narrowing the data spectrum, or both. Moving the upper bandedge of

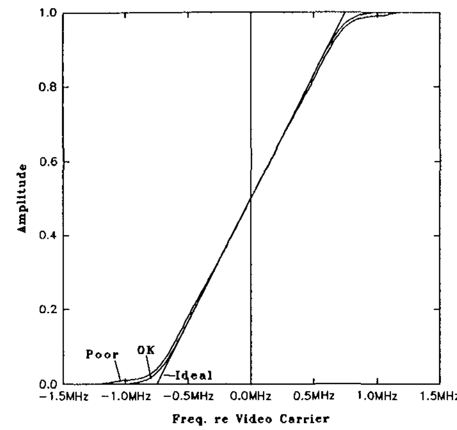

Figure 13 Receiver VSB Nyquist Filters

the data spectrum down by 100 kHz reduces the interference into the upper video picture by approximately 6 dB on problem receivers.

VI. Interference of Data into FM Sound

Our initial studies of this system concentrated on the interference of data into the BTSC sound signal. We found that with QPSK and alpha=0.7 filtering, the data carrier could be placed 350 kHz above the sound carrier. With alpha=0.3 filtering, the carrier could be moved down another 50 kHz so that it is only 300 kHz above the FM sound carrier. The mechanism for interference into FM sound is demodulation of the data signal as noise by the FM detector. The frequency of the demodulated noise is equal to the offset between the FM carrier and the data signal. Since the data signal is wideband, the demodulated noise is also wideband.

Fig. 14 shows the demodulated composite BTSC FM signal from DC to 200 kHz with no modulation on any of the BTSC channels. The 15.7 kHz pilot is clearly visible, along with L-R noise (produced by the BTSC compressor working at full gain) and the SAP carrier at 78.7 kHz. Overlaid is a plot with the data carrier turned on (using the parameters to be specified in section VIII). The data carrier causes a slightly higher noise level starting at about 90 kHz, and at 140 kHz the noise level rises abruptly. A higher noise level at these frequencies should be of no consequence in a properly designed mono, stereo,

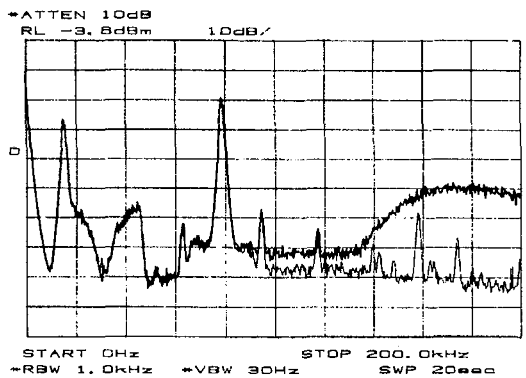

Figure 14 Spectrum from FM Demod, No Channels Modulated

or SAP decoder, and of only minor significance to a Pro channel decoder. A poorly designed decoder which inadvertently mixes components above 100 kHz down to baseband might suffer a slight degradation due to the addition of the data carrier, but this type of decoder would also suffer a premature noise degradation as CNR is reduced. Figure 15 shows the same plots, but with the Left and SAP channels fully modulated with a 400 Hz sinewave. This figure confirms that the noise induced into the composite baseband signal is similar whether the FM carrier is fully modulated or not. The presence of non-linear FM modulation does not appear cause the noise to spread down to lower audio frequencies.

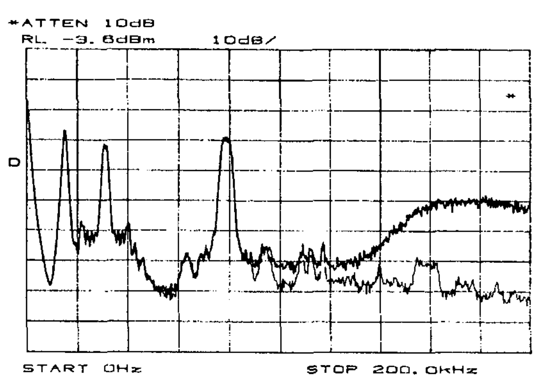

Figure 15 FM Demodulated Spectrum, All Channels Modulated

VII. Interference of Picture into Data

The system adopted in Scandinavia, which our first proposal was based on, has problems with interference of upper adjacent vision into data. The lower edge of the vestigial sideband energy extends low enough in frequency to be received by the data demodulator tuned to the lower channel. Their solution was to modify the spectral mask of allowed energy in the lower sideband. Basically, this involves changing out all VSB filters in cable modulators for new ones with the cutoff moved up approximately 200 kHz in frequency. Our original proposal would have required this change in modulator filtering as well.

When choosing new parameters to make the system more compatible with adjacent channel operation, it is very desirable to make the system operate with existing modulator filters. The steps taken to lower interference of data into picture will help to lower interference of picture into data.

VIII. A More Compatible System

Our original proposal of alpha=0.7 QPSK placed 350 kHz above the audio carrier at a -20 dB level is adequately compatible with BTSC audio, but could benefit from about 6 dB of improved compatibility with video. This could be achieved by dropping the carrier frequency by 100 kHz, but that would compromise compatibility with audio. Another way to achieve the 6 dB improvement is to narrow the carrier bandwidth as the frequency is dropped, keeping the lower data bandedge the same distance from the FM sound carrier. In theory, a 50 kHz drop in carrier frequency accompanied by a 100 kHz narrowing in bandwidth should meet our target. The narrowing could be done by staying with QPSK and changing the filtering from alpha=0.7 to alpha=0.3. The bandwidth will narrow by 0.4 x 250 = 100 kHz, and the carrier frequency can be dropped 50 kHz.

Unfortunately, when this is tried, the improvement in video interference is only about 3 dB. This is because the more tightly filtered data has a higher peak level, and the peaks are visible in the picture. The higher peak level has cost us 3 dB of the expected improvement. Changing from QPSK to QPRS, using the same RMS signal level, does not help, because QPRS inherently has a peak level which is 3 dB larger than the average level.

In making these comparisons, we have found it useful to normalize our signal levels to the nominal peak level, ignoring the overshoots. The eyes shown in Fig. 6 are normalized in this way, with a nominal peak level of $\pm$ 2 divisions. We measure the level of the modulated carrier with a static data pattern which produces a pure carrier modulated with this level. This is how we set the level in our original proposal. Using this method of measuring level, we find that we can meet our target by lowering the carrier frequency by 50 kHz and using either alpha=0.3 QPSK with a level of -23 dB, or QPRS with a level of -20 dB. These two signals will have comparable RMS levels, and using non-redundant error correction in the QPRS demodulator, similar performance in the presence of gaussian noise (the QPSK would actually be about 1.3 dB more rugged).

In order to choose which method to pursue, we have to look at the relative costs, and the performance in the presence of interfering signals. QPRS will require a more expensive decoder. Since the eye is multilevel, two comparators instead of one are required to convert from the analog waveform back to digital data. The level of the QPRS waveform at the comparator inputs is also critical, so a better AGC is required.

The data is subject to interference from both the FM audio carrier, and lower sideband visual information. The receive data filtering can adequately reject any interference from the FM carrier. Interference from video cannot be fully filtered out because it can be inband, depending on the VSB filter in the modulator. In order to improve rejection of video information, we have increased the order of the data receiver baseband low pass filtering from a 3 pole to a 5 pole filter. This change, along with the lowering of carrier frequency, allows our eyes to remain open even with the worst case interfering vision modulation present. With the worst case signal present our margin against error is reduced, but we can still operate error free.

If the QPRS and QPSK carriers were used with the identical carrier levels as defined above, the QPSK eye would start out twice as open as the QPRS eye. Since we can operate QPRS with 3 dB more level (as defined above, or at the same level on an RMS basis), the QPRS eye in the receiver will be 70% (-3 dB) the size of the QPSK eye. This smaller eye will be more susceptible to interference from lower sideband video information. Careful measurements show the -20 dB QPRS system suffers a 2 dB interference penalty compared to the -23 dB QPSK system. QPRS looses out on both circuit costs and ruggedness. Despite the attractively narrow spectrum of QPRS, QPSK is the superior solution. Therefore, should stay with QPSK.

The new QPSK carrier frequency will be 1.2 MHz below the vision carrier frequency. This value is very close to 1/3 of the NTSC chroma subcarrier frequency, which is 1.193182 MHz. This is a convenient value to choose, because it can allow future receiver circuits to use the chroma oscillator as a reference to demodulate the data, instead of a requiring a separate crystal oscillator to be locked to the incoming carrier.

The data rate may also be locked to video. We need approximately 512 k bits/sec of data to use a low cost digital audio coding method based on adaptive delta-modulation. It is desirable to use the same clock frequencies that are used in satellite implementations of this audio system such as B-MAC and HDB-MAC. Those systems place integral numbers of audio bits on horizontal lines and so have a direct relationship to video. It turns out that using 1/7 of the chroma frequency as a data clock gives us a total bit rate of 511.363 kHz. With the simplest conceivable multiplex structure, the audio clock rate will be 13 times the horizontal scan rate, which is identical to clock rate used by the BMAC systems. Using the chroma oscillator in the receiver as a data clock reference will save another crystal in the data timing recovery circuitry.

IX. The 1990 Proposal

The following seems to be the optimum set of parameters for an NTSC compatible digital carrier:

A. QPSK carrier with alpha=0.3 filtering.

B. Transmit filtering phase compensates for a specified receive filter.

C. Carrier frequency 1.193182 MHz (locked to 1/3chroma) below video carrier.

D. Carrier level -23 dB relative to peak vision carrier level.

E. Data rate 511.363 kHz (locked to 1/7 chroma).

IX. Conclusion

Minor modifications have been made to the NTSC compatible system described three years ago. We explored a number of methods of improving compatibility and settled on a new set of specifications. The new parameters improve compatibility, reduce receiver cost, and simplify transmission of the signal for both broadcasters an cable operators.

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