



ATSC Recommended Practice: Design Of Multiple Transmitter Networks

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ATSC Recommended Practice: **Design Of Multiple Transmitter Networks**

1 SCOPE AND INTRODUCTION

Many of the challenges of radio frequency transmission are the same regardless of whether the information carried is in analog or digital form. Because of the signal processing applied when the information carried is digital, however, there are techniques to overcome some of those challenges that are more applicable to digital signals than to analog signals. Among such techniques is the use of multiple transmitters in Single Frequency Networks (SFNs) and Multiple Frequency Networks (MFNs). In the past, SFNs have been considered mostly for applications in multi-carrier systems such as those using COFDM modulation. This Recommended Practice applies SFNs to the single-carrier 8-VSB system adopted by the ATSC and the FCC.¹ SFNs can be implemented with Digital On-Channel Repeaters (DOCRs), with Distributed Transmitters (DTxTs), with Distributed Translators (DTxRs), or with a combination of them. MFNs generally involve the use of translators. This Recommended Practice examines all three types of transmitters used in SFNs and MFNs and then concentrates on the design aspects of SFNs.

2 INFORMATIVE REFERENCES

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1. While this Recommended Practice is focused on the situation in the U.S., as regulated by the Federal Communications Commission (FCC), the fundamental application considerations discussed herein for single frequency networks are equally applicable in other countries and under other regulatory regimes.

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- [13] Hershberger, David: "Distributed Translator Systems for ATSC," NAB 2004 Broadcasting Engineering Conference *Proceedings*, pp.141–150, 2004.

3 DEFINITION OF TERMS

With respect to definition of terms, abbreviations, and units, the practice of the Institute of Electrical and Electronics Engineers (IEEE) as outlined in the Institute's published standards [1] are used. Where an abbreviation is not covered by IEEE practice or industry practice differs from IEEE practice, the abbreviation in question will be described in this document.

3.1 Compliance Notation

Descriptions of ATSC document types can be found in the ATSC Bylaws (B/2). Definitions of acceptable conformance terminology can be found in the ATSC Procedures for Technology and Standards Group Operation (B/3).

3.2 Treatment of Syntactic Elements

This document contains symbolic references to syntactic elements used in the audio, video, and transport coding subsystems. These references are typographically distinguished by the use of a different font (e.g., `restricted`), may contain the underscore character (e.g., `sequence_end_code`) and may consist of character strings that are not English words (e.g., `dynrng`).

4 CONSIDERATIONS FOR MULTIPLE TRANSMITTER NETWORKS

This section provides background information on key elements of multiple transmitter network design, specifically the challenges of RF transmission and the characteristics of digital signals. A comprehensive overview is also given of the basic architectures and characteristics of multiple transmitter networks, including the benefits and implementation challenges relating to such systems.

4.1 Challenges of RF Transmission

The economics and science of engineering a conventional television station coverage area are such that the last mile of coverage from a transmitter is far more expensive than the first mile of

coverage. Another way to think about this is to use the rule-of-thumb that, for a digital signal at UHF and at about 50 miles from a transmitter having an antenna 1200 feet above average terrain, it takes approximately an additional decibel of transmitter power to increase coverage by an additional mile. So, to cover an additional three miles requires doubling the transmitter power (+3 dB). To cover another three miles requires doubling the transmitter power again (now +6 dB), or four times the original power. Thus, increasing coverage with raw transmitter power can be expensive to accomplish.

Another set of challenges for RF transmission results from the lack of spectrum space that exists as a consequence of the effective doubling of the number of stations with the addition of a digital transmitter and channel for each full power analog broadcast station. At the same time the number of allotments has been doubled, the amount of spectrum available has been reduced with the withdrawal of those channels outside the “core spectrum.” There is also less space available on tall towers for adding antennas for high power transmitters given the number of additional transmitters that had to be built for digital operation; thus, “vertical real estate” is at a premium.

Interference issues add to the challenges of RF transmission. Another rule-of-thumb is that a digital UHF transmitter (as analyzed using the FCC F(50,10) curves) will cause co-channel interference to signals from another digital transmitter out to about three times the distance to which it can provide service (as analyzed using the FCC F(50,90) curves). Thus the co-channel interference zone of a transmitter will have a radius about three times the radius of its service zone. The result of this effect is that co-channel, high power transmitters must be widely spaced from each other in order to avoid interfering with one another.

When delivering signals to television receivers using set top antennas inside homes, the signal delivered to the vicinity of the homes must be sufficient to overcome the losses that occur when penetrating the buildings to reach the antennas. Building attenuation can be quite substantial, requiring that signals in the vicinity of a home be much stronger for indoor antenna reception than if an outdoor antenna were used. When factors are considered such as the likelihood that an indoor antenna is much closer to the ground than an outdoor antenna, thereby further reducing the signal level, a factor on the order of 40 dB of additional signal level is required to provide reliable reception using indoor antennas as opposed to rooftop antennas.

Further consideration of delivering signals to set top antennas reveals that many homes have metal of one sort or another included in the construction of their exterior walls. Examples of this are aluminum siding, aluminized wrappings used to improve insulation, and metal lath used to hold stucco on walls. Such metal enclosures turn homes into virtual resonant cavities, with windows acting as ports that allow signals in the outside environment to excite the cavities. The result of this arrangement will be quite significant standing waves within a home, leading to areas having usable signal levels and others having very low signal levels. Putting aside for the moment the matter of multipath, whether a receiver will work inside a home will depend upon its placement relative to the standing waves and the signal level that it receives as a result.

One more factor that challenges RF transmission is the terrain. Especially at UHF, terrain can block signals from transmitters that do not have line-of-sight to receivers. This can make it very difficult to deliver adequate signal levels to locations that are terrain-blocked (i.e., “propagationally challenged”), even when those locations are close enough to transmitters that they should otherwise receive more than adequate signal levels.

4.2 Characteristics of Digital Signals

Digital signals have a set of characteristics that make operation of the RF transmission environment even more complex. They start with all of the general characteristics that impact transmission of analog signals since, in reality, they are analog signals that carry information in such a way that it can be interpreted to recover information in the form of bits. Thus, they suffer from all the degradations, distortions, and impairments that impact analog signals. These can include such linear distortions as amplitude variations across the frequency spectrum of the channel, envelope delay distortion, and multipath. They also can include such non-linear distortions as AM-AM and AM-PM (equivalent to differential gain and differential phase in analog signals).

Because the various distortions, if uncorrected, can make it impossible to recover the data from the signal, various forms of digital signal processing are applied to the signals to enable detection of the data under a wide variety of conditions. Non-linear distortions are removed, to the extent possible, at the transmitter. Linear distortions of the type caused by the transmission channel are treated, at least conceptually, by developing a model of the channel and then applying its inverse to the signal. The determination of the channel model and the creation of a filter having characteristics inverse to those of the channel are the functions of an adaptive equalizer, one of which is included in practically every digital receiver. The capabilities of its adaptive equalizer determine how much linear channel distortion each receiver can withstand while still accurately receiving the signal and recovering the data.

The characteristics of adaptive equalizers that are significant in determining their ability to correct for various types of channel impairments include the time offset of the echoes that can be handled — both pre-cursor and post-cursor, the strength of the echoes that can be managed relative to the main signal at any given delay from the main signal, the amount of Doppler shift in the echoes that can be dealt with, and the rapidity with which the channel characteristics can change and still be tracked and corrected. It is also important to know the relative signal level at which the receiver treats an interfering echo signal as just noise and does not need the adaptive equalizer to correct it. Some adaptive equalizers also are capable of combining the energy from echoes with that of the main signal, thereby benefiting from the echoes and adding power toward meeting the threshold at which the signal can be received.

Even with the application of adaptive equalizers, it is the nature of digital transmission that noise in the channel causes errors in the recovery of the data bits from the signal. To overcome this issue, error correction coding (ECC) is routinely used in digital transmission systems. In the ATSC 8-VSB system, two types of ECC are used: Reed Solomon coding is applied to blocks of data that are transmitted together as data segments, and trellis coding is applied to the data symbols as they are transmitted. Error correction coding of these types makes it possible to accurately recover the data after much greater channel impairments and at much lower signal levels than would be possible without ECC. As the signal level decreases, however, the ECC decoders in receivers abruptly lose the ability to correct the increasing number of errors, leading to sudden failure of reception. This is called the “cliff effect.”

One more aspect of digital transmission that must be considered is the set of planning factors used to determine the conditions under which reception should be possible. Planning factors include such characteristics of the receiving system as the antenna gain, transmission line losses, impedance mismatch losses, receiver noise figure, and the threshold signal-to-noise ratio (S/N) of the receiver. Recent work in the ATSC has exposed the possibility that the planning factors

originally used to establish channel allotments, while adequate for providing paired allotments for all stations, may require adjustment by as much as 10 dB in order to more accurately predict coverage. This means that up to 10 dB more power may be needed at the receiver to reliably receive the signals. Multiple transmitter networks can help address such a requirement.

4.3 Multiple Transmitter Networks

Many of the challenges of RF transmission, especially as they apply to digital transmission, can be addressed by using multiple transmitters to cover a service area. Because of the limitations in the spectrum available, many systems based on the use of multiple transmitters must operate those transmitters all on the same frequency, hence the appellation Single Frequency Network. At the same time, use of SFNs leads to a range of additional complications that must be addressed in the design of the network.

SFNs for single-carrier signals such as 8-VSB become possible because of the presence of adaptive equalizers in receivers. When signals from multiple transmitters arrive at a receiver, under the right conditions, the adaptive equalizer in that receiver can treat the several signals as echoes of one another and extract the data they carry. The conditions are controlled by the capabilities of the adaptive equalizer and will become less stringent as the technology of adaptive equalizers improves over time. This will allow more flexible designs of SFNs while maintaining a particular level of reliability of signal delivery.

4.3.1 Benefits of Multiple Transmitters

A number of benefits may accrue to the use of multiple transmitters to cover a service area. Among these can be the ability to obtain more uniform signal levels throughout the area being served and the maintenance of higher average signal levels over that area. These results come from the fact that the average distance from any point within the service area to a transmitter is reduced. Reducing the distance also reduces the variability of the signal level with location and time and thereby reduces the required fade margin needed to maintain any particular level of reliability of service. These reductions, in turn, permit operation with less overall effective radiated power (ERP) and/or antenna height.

When transmitters can be operated at lower power levels and/or elevations, the interference they cause to their neighbors is reduced. Using multiple transmitters allows a station to provide significantly higher signal levels near the edge of its service area without causing the level of interference to its neighbor that would arise if the same signal levels were delivered from a single, central transmitter. The interference reductions come from the significantly smaller interference zones that surround transmitters that use relatively lower power and/or antenna heights.

With the use of multiple transmitters comes the ability to overcome terrain limitations by filling in areas that would otherwise receive insufficient signal level. When the terrain limitations are caused by obstructions that isolate an area from another (perhaps the main) transmitter, advantage may be taken of the obstructions in the design of the network. The obstructions can serve to help isolate signals from different transmitters within the network, making it easier to control interference between the network's transmitters, as will be described below. When terrain obstructions are used in this way, it may be possible to place transmitters farther apart than if such obstructions were not utilized for isolation.

Where homes are illuminated by sufficiently strong signals from two or more transmitters, it may be possible to take advantage of the multiple signals to provide more reliable indoor reception. When a single transmitter is used, standing waves within a home sheathed in metal

likely will result in areas within that home having signal levels too low to use. Signals arriving from different directions will enter the resonant cavity of the home through different ports (windows) and set up standing waves in different places. The result often may be that areas within the home receiving low signal levels from one transmitter will receive adequate signal levels from another transmitter, thereby making reliable reception possible in many more places within the home.

4.3.2 Limitations of SFNs

While they have the potential to solve or at least ameliorate many of the challenges of RF transmission, SFNs also have limitations of their own. Foremost among these is the fact that there will be interference between the signals from the several transmitters in a network. This “system-internal” interference must be managed so as to bring it within the range of capabilities of the adaptive equalizers of the largest number of receivers possible. Where the interference falls outside the range that can be handled by a given adaptive equalizer, other measures, such as the use of an outdoor directional antenna, must be applied.

The characteristics of adaptive equalizers that are important for single frequency network design are the ability to deal with multiple signals (echoes) with equal signal levels, the length of echo delay time before and after the main (strongest) signal that can be handled by the equalizer (or the total delay spread of echoes that can be handled when an equalizer design does not treat one of the signals as the main signal), and the interfering signal level below which the adaptive equalizer is not needed because the interference is too low in level to prevent reception.

The last of these characteristics, which is somewhat lower in level than the co-channel interference threshold, determines the areas where the performance of the adaptive equalizer matters. In places with echo interference below the point at which the adaptive equalizer needs to correct the signal, it is not necessary in design of the SFN to consider the differential arrival times of the signals from the various transmitters. In places with echo interference above that threshold, the arrival times of the signals matter if the adaptive equalizer is to be able to correct for the echo interference. Differential delays between signals above the echo interference threshold must fall within the time window that the adaptive equalizer can correct if the signals are to be received.

Current receiver designs have a fixed time window inside which echoes can be equalized. The amplitudes of correctable echoes also are a function of their time displacement from the main signal. The closer together the signals are in time, the closer they can be in amplitude. The further apart they are in time, the lower in level the echoes must be for the equalizer to work. These relationships are improving dramatically in newer receiver front-end designs, and they can be expected to continue improving at least over the next several generations of designs. As they improve, limitations on SFN designs will be reduced.

The handling of signals with equal signal level echoes is a missing capability in early receiver front-end designs, but it is now recognized as necessary for receivers to work in many situations that occur naturally, without even considering their generation in SFNs. The reason for this is that any time there is no direct path or line-of-sight from the transmitter to the receiver, the receiver will receive all of its input energy from reflections of one sort or another. When this happens, there may be a number of signals (echoes) arriving at the receiver that are about equal in amplitude, and they may vary over time, with the strongest one changing from time-to-time. This is called a Rayleigh channel when it occurs, and it is now recognized that Rayleigh channels are more prevalent than once thought. For example, they often exist in city canyons and mid-rise areas, they exist behind hills, and so on. They also exist in many indoor situations. If receivers are

to deal with these cases, adaptive equalizers will have to be designed to handle them. Thus, SFNs will be able to take advantage of receiver capabilities that are needed in natural circumstances.

Radio frequency signals travel at a speed of about 3/16 mile per microsecond. Another way to express the same relationship is that radio frequency signals travel a mile in about 5-1/3 microseconds. If a pair of transmitters emits the same signal simultaneously and a receiver is located equidistant from the two transmitters, the signals will arrive at the receiver simultaneously. If the receiver is not equidistant from the transmitters, the arrival times at the receiver of the signals from the two transmitters will differ by 5-1/3 microseconds for each mile of difference in path length. In designing the network, the determination of the sizes of the cells and the related spacing of the transmitters will be dependent on this relationship between time and distance and on the delay spread capability of the receiver adaptive equalizer.

Because receivers have limited delay-spread capability, there is a corresponding limit on the sizes of cells and spacing of transmitters in SFNs. As receiver front-end technology improves over time, this limitation can be expected to be relaxed. As the limitation on cell sizes is relaxed and cells become larger, it can be helpful to network design to adjust the relative emission times of the transmitters in the network. This allows putting the locus of equidistant points from various transmitters where needed to maximize the audience reached and to minimize internal interference within the network. When such time offsets are used, it becomes desirable to be able to measure the arrival times at receiving locations of the signals from the transmitters in the network. Such measurements can be difficult since the transmitters are intentionally transmitting exactly the same signals in order to allow receivers to treat them as echoes of one another. Somehow the transmitters have to be differentiated from one another if their respective contributions to the aggregate signal received at any location are to be determined. To aid in the identification of individual transmitters in a network, a buried spread spectrum pseudorandom “RF Watermark” signal is included in ATSC A/110 [3], as described in detail below.

4.3.3 Single Frequency Networks

Table 4.1 lists the most common forms of distributed transmitter networks and some of their primary characteristics.

Table 4.1 Comparison of Different Distributed Transmitter Networks

SFN Configurations	Distributed Transmitters	Digital On Channel Repeaters	Distributed Translators
Complexity/Cost	High	Low	Low
Requirement for Feeder Link	Yes	No	No
Delay Adjustment Capability	Yes	No	Yes
Output Power Level	No Limit	Low/Moderate	No Limit
Number of DTV RF Channels Required	One	One	At Least Two
Recommended Implementation	Large Area SFN	Cover Small Area: Gap Filler, and Coverage Extender	Large Area SFN, Gap Filler, and Coverage Extender

4.3.3.1 Digital On-Channel Repeaters

In block diagram form, Digital On-Channel Repeaters (DOCRs) look just like translators or boosters. (See Figure 4.1.) They comprise a receiving antenna, a receiver, some signal processing, a transmitter, and a transmitting antenna. They are closest in nature to boosters in that they receive

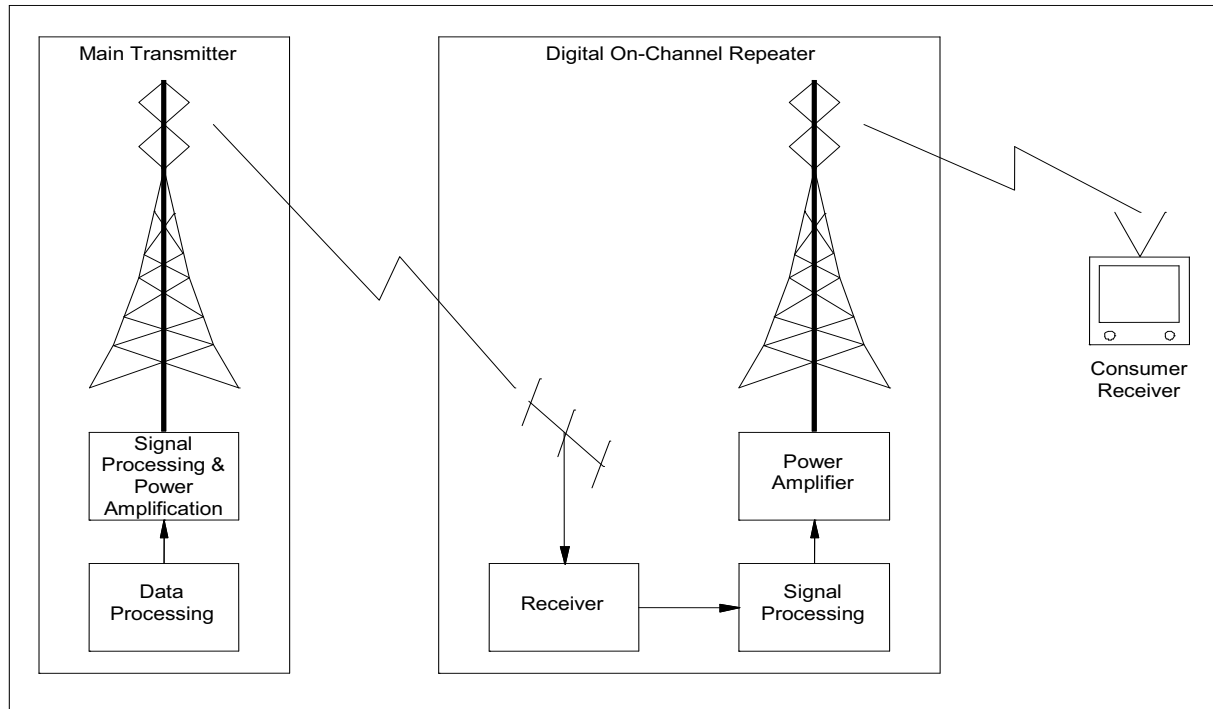


Figure 4.1 Digital on-channel repeater (DOCR) generic block diagram.

the off-the-air DTV signal, process it, and retransmit it on the same frequency. This greatly simplifies the portion of a Single Frequency Network that employs them, but it also leads to limitations in how they can be applied. For example, due to the internal signal processing and filtering delays of a DOCR, the signal from the main transmitter will arrive at the DOCR coverage area first, which acts as a pre-echo, relative to the output signal from the DOCR. This type of multipath distortion is very harmful to reception by ATSC legacy receivers. With the anticipated performance improvements of newer receivers, this problem is expected to be reduced or eliminated in the future. The main advantage for the DOCR is its simplicity and low cost. With a DOCR, there is no need for a separate Studio-to-Transmitter Link (STL).

There are several configurations that can be used in DOCRs. They are shown in Figures 4.2a through 4.2d, and can be designated as:

- RF Processing DOCR (Figure 4.2a)
- IF Processing DOCR (Figure 4.2b)
- Baseband Demodulation and Equalization DOCR (Figure 4.2c)
- Baseband Decoding and Re-generation DOCR (Figure 4.2d)

Figure 4.2 DOCR configurations (*see next page*).

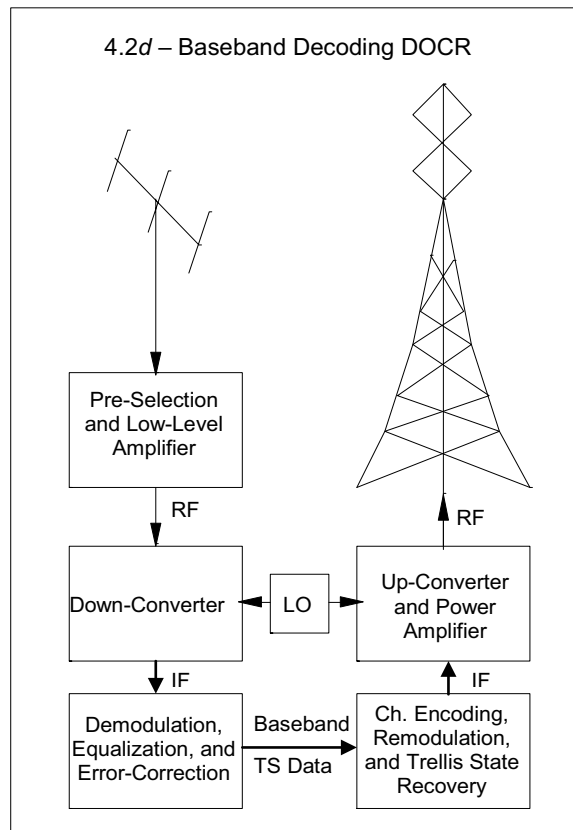
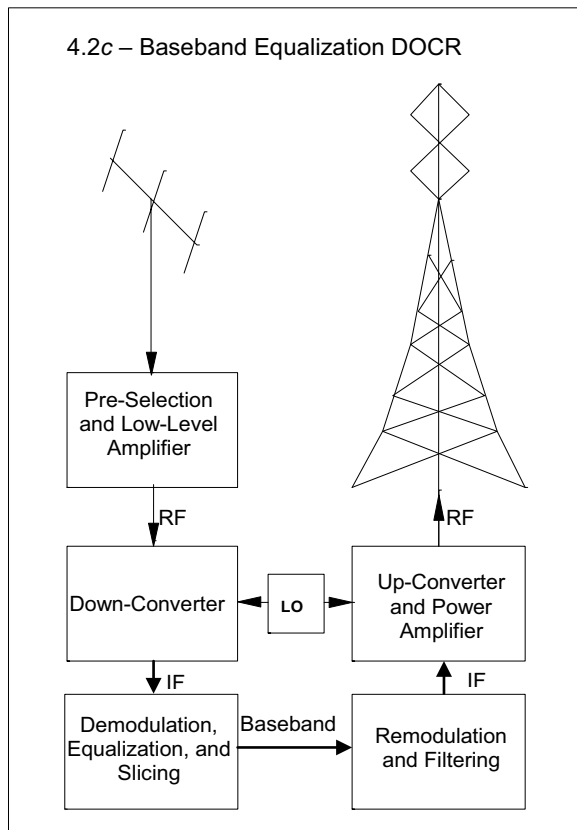
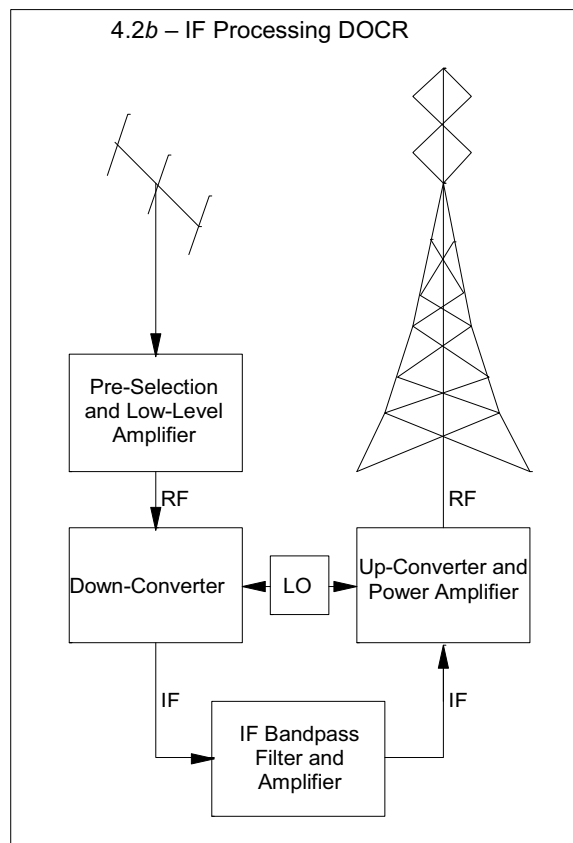
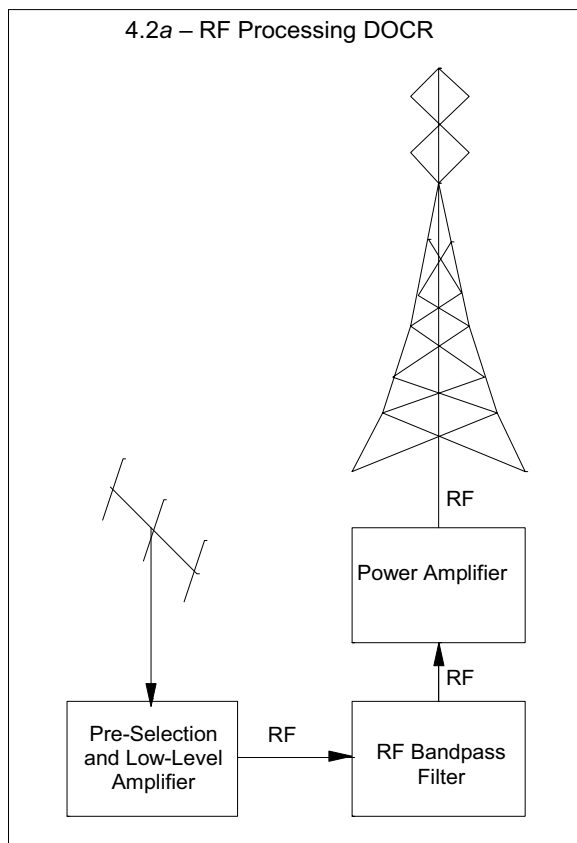


Figure 4.2*a* shows the simplest form of DOCR, the RF Processing DOCR, in which the receiver comprises a preselector and low-level amplifier, the signal processing comprises an RF bandpass filter at the channel of operation, and the transmitter comprises a power amplifier. There is no frequency translation of any sort in this arrangement. It has the shortest processing delay of any of the DOCR configurations — usually a fraction of a microsecond. This configuration, however, has very limited first adjacent channel interference rejection capability. Such a configuration can result in the generation of intermodulation products in the amplifier and in degradation of the re-transmitted signal quality. Meanwhile, due to limited isolation (i.e., too much coupling) between transmitting and receiving antennas, the re-transmitted signal could loop back and re-enter the receiving antenna. This can also degrade the signal quality, causing spectrum ripple and other distortions. The only way to limit or avoid the signal loopback in this type of DOCR is to increase antenna isolation, which is determined by the site environment, or to limit the DOCR output power. Usually, the RF Processing DOCR transmitter output power is less than 10 W, resulting in an effective radiated power (ERP) on the order of several dozen watts.

Conversion to an intermediate frequency (IF) for signal processing is the principal feature that differentiates Figure 4.2*b*. In this arrangement, a local oscillator and mixer are used to convert the incoming signal to the IF frequency, where it can be more easily amplified and filtered. The same local oscillator used for the downconversion to IF in the receiver can be used for upconversion in the transmitter, resulting in the signal being returned to precisely the same frequency at which it originated (with some amount of the local oscillator phase noise added to the signal). The delay time through the IF Processing DOCR will be decided mostly by the IF filter implemented. A SAW filter can have much sharper passband edges, better control of envelope delay, greater attenuation in the stopband, and generally more repeatable characteristics than most other kinds of filters. Its transit delay time for the signal passing through it can be in the order of 1–2 microseconds, so the delay through an IF Processing DOCR, Figure 4.2*b*, will be from a fraction of a microsecond to about 2 microseconds — somewhat longer than the RF Processing DOCR. The IF Processing DOCR has better first adjacent channel interference rejection capability than does the RF Processing DOCR, but it retains the signal loopback problem, which limits its output power.

Figure 4.2*d* shows a receiver that demodulates the incoming signal all the way to a digital baseband signal in which forward error correction (FEC) can be applied. This restores the bit stream to perfect condition, correcting all errors and eliminating all effects of the analog channel through which the signal passed in reaching the DOCR. The bit stream then is transmitted, starting with formation of the bit stream into the symbols in an exciter, just as in a normal transmitter. If no special steps are taken to set the correct trellis encoder states, the output of the DOCR of Figure 4.2*d* would be incoherent with respect to its input. This would result in the signal from such a repeater acting like noise when interfering with the signal from another transmitter in the network rather than acting like an echo of the signal from that other transmitter. Thus, additional data processing is required to establish the correct trellis states for retransmission when such a DOCR is used. It should also be noted that this form of DOCR has a very long delay through it, measured in milliseconds, mostly caused by the deinterleaving process. This delay is well outside the ATSC receiver equalization range. Although, by regenerating the DTV signal, it totally eliminates the adjacent channel interference and signal loopback problems, this type of DOCR has very little practical use, unless the intended DOCR coverage area is totally isolated from the main DTV transmitter.

A more practical intermediate method is the Baseband Equalization DOCR, or EDOCR, as shown in Figure 4.2*c*. It fits between the techniques of Figures 4.2*a* or 4.2*b* and 4.2*d*. This type of DOCR demodulates the received signal and applies adaptive equalization in order to reduce or eliminate adjacent channel interference, signal loopback, and multipath distortion occurring in the path from the main transmitter to the EDOCR. In determining the correct 3-bit symbol levels, it also carries out symbol level slicing or trellis decoding, which can achieve several dB of noise reduction and minimizes the impact of channel distortions. The baseband output of the equalizer and slicer is re-modulated, filtered, frequency shifted and amplified for re-transmission. The delay of the baseband processing is in the order of a few dozen VSB symbol times. The total EDOCR internal delay is in the order of a few microseconds once the time delays of the adaptive equalizer and the pulse shaping (root raised cosine) filters (one each for receive and transmit) are taken into account. This amount of delay could have an impact on ATSC legacy receivers. The Baseband Equalization DOCR allows retransmission of a clean signal without the lengthy delays inherent in the Baseband Decode/Regeneration method, of Figure 4.2*d*. It can also transmit at higher power than that of the RF and IF Processing DOCRs. The shortcoming of this method is that, since there is not complete error correction, any errors that occur in interpretation of the received data will be built into the retransmitted signal. This makes it important in designing the EDOCR installation to include sufficient receiving antenna gain and received signal level that errors are minimized in the absence of error correction. If a fairly clean received signal cannot be obtained, it may be better to use the RF or IF Processing DOCR (Figure 4.2*a* or 4.2*b*) and allow viewers' receivers to apply full error correction to the relayed signal rather than retransmitting processed signals that contain data errors.

All of the forms of DOCR described have a number of limitations in common. Most of the limitations arise from the facts that a DOCR receives and re-transmits on the same frequency and that obtaining good isolation between transmitting and receiving antennas is a very difficult proposition. The result is coupling from the DOCR output back to its input. This coupling leads to feedback around the amplifiers in the DOCR. In the limit, such feedback can result in oscillation in the cases of the RF and IF Processing DOCR. Short of oscillation, it can result in signal distortions in amplitude and group delay ripple similar to those suffered in a propagation channel. These two designs will also suffer from the accumulation of noise along the cascade of transmitters and propagation channels from the signal source to the ultimate receiver. The application of adaptive equalizers to the feedback path around DOCRs holds some promise to mitigate the distortions, but it cannot help with the noise accumulation. The Baseband Decode/Regeneration DOCR, Figure 4.2*d*, will eliminate the noise and distortion accumulation, but at the cost of a time delay that is likely to be unacceptable for network design.

The feedback around a DOCR puts a limitation on the power that can be transmitted by such a device. A margin must be provided to keep the system well below the level of oscillation, and the point of oscillation will be determined by the isolation between the transmitting and receiving antennas. All of this tends to make high power level operation of DOCRs problematic.

Similarly, the time delay through a DOCR is significant for network design. As one goes from design-to-design from Figure 4.2*a* to 4.2*b*, 4.2*c*, and then 4.2*d*, the time delay gets longer. The time delay determines over what area the combination of signals from the transmitters in the network stay within the capability of the receiver adaptive equalizer to correct the apparent echoes caused by receiving signals from multiple transmitters. The geometry between the source transmitter, the DOCR, and the receiver determine the delay spread actually seen by the receiver. To this must be added the delay of the DOCR. Additional delay in the DOCR can only push the

delay spread in the wrong direction (extending pre-echo), further limiting the area in which a receiver having a given adaptive equalizer capability will find the delay spread within its ability to correct.

DOCRs offer the lowest cost per cell in a single frequency network, but they are limited in the areas they can cover by the power and time delay limitations described above. Thus, they require more cells to cover a large area than other solutions to be described. Nevertheless, they have an important role to play in SFNs within their limitations. Indeed, they are likely to be used in combination with the more flexible solution offered by the Distributed Transmission Network. The major advantage of a DOCR is its spectrum efficiency and cost effectiveness. It does not need an STL. The pros and cons of different DOCR configurations are summarized in Table 4.2.

Table 4.2 Performance Comparison of Different DOCR

DOCR Configurations	RF Processing DOCR	IF Processing DOCR	Baseband Equalization DOCR	Baseband Decode/ Regeneration DOCR
Complexity/Cost	Very Low	Low	High	High
DOCR Internal Delay (Extending Pre-echo)	Few tenths μ s, Very Small	Around 1 – 2 μ s, Small	Several μ s, Medium	mS Level, Beyond Equalizer Range
Input Adjacent Channel Suppression Capability	Very Weak	Medium	High	High
Output Power Level	Low	Low	Moderate	Moderate
Multihop Noise/Error Accumulation	Yes	Yes	Maybe	No

4.3.3.2 Distributed Transmission Systems

Distributed Transmission Networks differ from DOCRs in that each transmitter in the network is fed over a link separate from the on-air signal delivered to viewers (although this may be over a different broadcast channel in the case of distributed translators). This separation of the delivery and emission channels provides complete flexibility in the design of individual transmitters and the network, subject only to the limitations of consumer receiver adaptive equalizer capabilities. From the standpoint of the network and its transmitters alone, any power level and any relative emission timing are possible. The important requirement is to assure that all transmitters in the network emit the same symbols for the same data delivered to their inputs and at the same time (plus whatever time offsets may be designed into the network).

There are a number of ways in which signals in a DTxN could be delivered to the several transmitters. These include methods in which fully modulated signals are delivered to the various transmitter locations and either up-converted or used to control the re-modulation of the data onto a new signal, a method in which data representative of the symbols to be broadcast are distributed to the transmitters, and a method in which standard MPEG-2 Transport Stream packets are delivered to the transmitters together with the necessary information to allow the various transmitters to emit their signals in synchronization with one another. It is the latter method that has been selected for use in ATSC A/110 [3].

In the method documented in A/110, the data streams delivered to the transmitters are the same as now delivered over standard STLs for the single transmitters currently in use, with information added to those streams to allow synchronizing the transmitters. This method utilizes a very small portion of the capacity of the channel but allows continued use of the entire existing infrastructure designed and built around the 19.39 Mb/s data rate. This technique permits complete flexibility in setting the power levels and relative emission timing of the transmitters in a

network while assuring that they emit the same symbols for the same data inputs. While originally intended for use in SFNs, the selected method also permits extension to MFNs, using a second broadcast channel as an STL to deliver the data stream to multiple distributed translators that themselves operate in an SFN. Various combinations of distributed transmitters and distributed translators are possible, and, in some cases, whether a given configuration constitutes an SFN or an MFN will depend only upon whether there are viewers in a position to be able to receive the signals that are also relaying the data stream to successive transmitters in the network.

4.3.4 Multiple Frequency Networks

A multiple frequency network (MFN) uses more than one channel for transmission. In the purest case, for N transmitters, N channels are used. But where DTx technology is being applied, channels may be shared among a number of transmitters. For N transmitters, the number of channels used will be less than N . Some of the transmitters will be synchronized, operating on the same channel. In this situation, the network is actually a hybrid of multiple frequency and single frequency techniques. This classification (hybrid, multiple frequency, DTx networks, where some transmitters are synchronized) is the subject of this section.

4.3.4.1 Translators

A translator is part of a multiple frequency network. It receives an off-air signal on one channel and retransmits it on another channel.

4.3.4.2 Distributed Translators

Even in some relatively unpopulated areas, especially where NTSC translator systems are already deployed, there are not enough additional channels to accommodate traditional translator networks for ATSC signals during the transition phase. In these situations, use of distributed translators allows ATSC translator systems to be built using fewer channels.

A distributed translator system applies DTx technology to create a network of synchronized transmitters on one channel, which retransmits a signal received off-the-air from a main transmitter, distributed transmitter, another translator, or a translator network. The advantages of a distributed translator system over conventional translators include: 1) signal regeneration, and 2) conservation of spectrum.

The distributed transmission system initially was designed to use STLs to convey MPEG-2 Transport Stream streams to slave transmitters in a network. This certainly could be done for distributed translator systems, but where an off-air signal is available, use of STLs would be costly and redundant. The advantage of a distributed translator system over a distributed transmission system is that STLs are not required — the signal may be taken off the air.

Use of off-air signals for distributed translators introduces some problems that can be overcome using techniques described here.

The DTxA makes three kinds of changes to a MPEG-2 Transport Stream signal. These are:

- Insertion of data into the MPEG header error flag bit (the field rate side channel)
- Periodic inversion of the MPEG sync byte (cadence sync)
- Insertion of DTxPs

The simplest example of a distributed translator system would be one where a single main transmitter is used as an “STL” — it would operate as an ordinary ATSC transmitter rather than as

a slave. The DTxP would pass through the transmitter unmodified and be broadcast. The trellis codes would remain intact because the main transmitter is not a slave.

Distributed translators would receive the off-air signal and retransmit it on another channel. The information in the DTxP enables proper timing and synchronization.

The field rate side channel and the cadence sync, however, are necessarily lost in this situation. The field rate side channel is lost because the MPEG header error flag bit must be restored to zero before transmission; otherwise consumer receivers would reject MPEG packets incorrectly indicated as errored.

Cadence sync, which is a periodic inversion of the MPEG sync byte, also is lost because MPEG sync is not transmitted. (It is replaced by segment sync.)

The field rate side channel is used mainly for E-VSB applications, where the field sync reserved bits may change as often as once per field. For non E-VSB transmissions, the field sync reserved bits can be set to their default values, rather than using field rate side channel data to set them. For E-VSB applications, a special receiver would be required that recovers the field sync reserved bits and passes them on to the slave transmitters.

When the DTx-modified MPEG-2 Transport Stream bitstream is transmitted as an ATSC signal, the periodic inversion of the MPEG sync byte is lost. The periodic inversion is used to identify the position where ATSC frame sync is to be inserted. Fortunately, there is another way to determine the position of frame sync. The DTxP contains a pointer to the frame sync position. A value exists within the DTxP that identifies its packet number with respect to frame sync. The packet number value is an integer between 0 and 623, with frame sync appearing before packet 0. This value, although it will generally be received less often than cadence sync, provides the necessary frame sync phasing information.

For single-tier distributed translator systems, the DTxP is processed in the DTxA and the slave DTxRs as it would be in any distributed transmission system. When the DTxP is retransmitted by the distributed translators, however, the trellis code data are necessarily obliterated.

In this situation, a second tier of translators repeating the signal from the first tier would be impossible for two reasons. First, the trellis codes would have been removed from the DTxP in the first tier of translators. Second, a subsequent tier of translators would require additional time delay prior to emission to allow for demodulation, deinterleaving, error correction, reinterleaving, remodulation, etc.

Both of these problems can be solved if an additional layer of DTxPs is inserted for each tier of translators. Each tier of translators would process only the DTxPs addressed to it, and would allow all of the others to pass unmodified.

To accomplish this, a way of associating a layer of DTxPs with a tier of translators has been developed. The OMP (operations and maintenance) type byte is used for this purpose. For a conventional DTx system, the OMP value is 0x00. In a situation with a single, main transmitter, for the first tier of distributed translators, the OMP value would continue to be 0x00. For the second tier of distributed translators, the OMP byte would be set to 0x01, etc., up to a maximum of 0x1F. This allows up to 32 cascaded tiers of distributed translators. 32 cascaded translators could be considered an impracticably large value for analog translators; for ATSC, however, there will be data regeneration at each tier. So long as there are no uncorrectable transmission errors, the signal quality at the last tier will be just as good as it is from the main transmitter.

Formation of the tiered DTxPs for distributed translators must occur in a certain order. It might appear that a simple cascade of multiple DTxAs would form the correct MPEG-2 Transport

Stream output signal, but this is not the case. In a simple cascade, the first DTxA would control the last tier of DTxRs, the second DTxA would control the next to the last tier of DTxRs, etc. The DTxPs for the first tier would be formed by the last DTxA in the chain. Obviously, these DTxPs formed by the last DTxA would not be present in the first DTxA, and they would not pass through the first DTxA's coding model. Therefore, they could not pass through the coding model in the last tier of distributed translators, either. These packets would have to be restored to the blank DTxP placeholder state before being transmitted. By removing all data from these DTxPs, DTx network information would be denied to test and measurement equipment in the field.

Consequently, a different ordering of packet formation is required for multiple-tier DTxR operation to allow all of the tiers of DTxRs to transmit all of the DTxPs. In a conventional DTxA, the packet processing consists of the formation of the packets, less trellis codes and Reed-Solomon FEC (denoted as process "P"), and channel coding and insertion of trellis codes and Reed-Solomon FEC into the packet (denoted as process "T"). Referring to these steps as P_n and T_n for tier n , a cascade of three conventional DTxAs would perform the processing in the order $P_2, T_2, P_1, T_1, P_0, T_0$. The processing for the outer tier 2 must be done first, the middle tier 1 is second, and the innermost tier 0 is last. But for multiple-tier DTxR operation, all of the packets for all of the tiers (less trellis codes and Reed-Solomon FEC) must be formed before the first coding model (which corresponds to the last tier of DTxRs). That ordering allows all packets to pass through each tier's coding model and subsequently to be transmitted. Using the same notation as above, the re-ordered processing of the MPEG-2 Transport Stream bitstream becomes $P_2, P_1, P_0, T_2, T_1, T_0$. For a system operating with tiers 0 through n ($n+1$ tiers), the processing order is $P_n, P_{n-1}, \dots, P_0, T_n, T_{n-1}, \dots, T_0$. (Although the processing order of the channel coding/trellis code operations is critical, the order in which the packets are formed is not, except that all packets must be formed before the channel coding/trellis code operations. In other words, the sequence $P_n, P_{n-1}, \dots, P_0, T_n, T_{n-1}, \dots, T_0$ above is equivalent to $P_0, P_1, \dots, P_n, T_n, T_{n-1}, \dots, T_0$). Formation of the DTxPs in this order will ensure that all of the DTxPs are transmitted by all of the transmitters in the network.

When setting up the system parameters of a DTxR system, each tier of the system must be delayed at least approximately 10 milliseconds with respect to its predecessor. This delay allows time for receiver decoding and transmitter encoding delays at the translators (a large portion of which is due to interleaving).

4.3.4.3 Baseband Equalization Distributed Translators EDTxR

If over-the-air signal received at a translator site is fairly clean with sufficient level, it may not be necessary to demodulate and decode the signal all the way down to a baseband digital stream to apply forward error correction (FEC). Bringing the signal down to baseband (but not decoding it) and applying adaptive equalization to compensate for channel distortions, and then retransmitting on another channel may be enough to achieve a firmly acceptable signal in the coverage area of the translator. Since no decoding and re-encoding of the signal is involved in this process, the trellis code modulated (TCM) symbols are not altered in the translator and the pattern of the symbol levels remains the same as that in the input signal.

When two or more of such translators receive their input signal from the same source, they emit the same symbol patterns (with a time offset that can be tailored for each translator) without requiring any additional means such as DTxP. Under such conditions, the EDTxRs do not have any TCM encoder ambiguity problem and if the output frequency and clock of the translators

(transmitting on the same channel) are synchronized, they can form a distributed translator network.

The structure of EDTxR is basically the same as that of EDOCR (explained in Section 4.3.3.1 and shown in Figure 4.2c). However, difference in the input and output channel frequencies of EDTxR results in less restriction in its design considerations. This in turn leads to a number of differences in the implementation and in the characteristics of EDTxR as compared to EDOCR. Such differences include:

- 1) As there is no RF co-channel signal loopback from output to input of a translator, the output power limitations of EDOCR do not apply to EDTxR.
- 2) Because there is no co-channel reception in the overlapping coverage areas of a main transmitter and a translator, the time delay introduced by the translator to the main transmitter signal is not critical. Therefore, there is less restrictions in using more appropriate filtering and signal processing (encountering more delays) inside EDTxR. Each translator, however, should be capable of providing an adjustable additional delay to the signal to enable controlling and adjusting the emission timing of the translators with respect to each other.
- 3) In EDOCR, the output frequency of the repeater should be synchronized with its input frequency and any offset between the two frequencies can be quite harmful to the receiver. In EDTxR, however, the output frequency of different translators should be synchronized with each other and the offset with the input frequency is not a matter of concern. This can lead to a more specific synchronization procedure for EDTxR that may be different from that of EDOCR.

Multi-tier operation of EDTxR is possible as far as the accumulated errors in the channels through which the signal has passed to reach the last translator have not gone beyond an acceptable limit. The output frequency of the translators in each tier, however, should be precisely synchronized.

4.3.5 A Balancing of Trade-Offs

A generalized DTx system may include different combinations of DTxTs, DTxRs, and DOCRs. Nothing inherently prohibits using one in the presence of another. System architects may tailor the deployment of the various technologies to the particular system being designed, considering such issues as power, cost, and terrain shielding.

For example, where low power is adequate, particularly where there is terrain shielding, a DOCR may be the most cost-effective solution. If higher power is necessary than a DTxT may be better. DTxTs also offer control of delay. Where the cost of an STL would be too high, a DTxR system may be best, assuming that an additional channel is available.

5 APPLICATIONS OF SINGLE FREQUENCY NETWORKS

This section describes three main techniques for extending DTV coverage — Digital On-Channel Repeaters (DOCRs), Distributed Transmitters (DTxTs), and Distributed Translators (DTxRs). All of these approaches are intended to be complementary to one another and may be used together or individually as tools for the DTV broadcast system designer. Each technique comes with advantages and disadvantages, which are outlined below.

5.1 Digital On-Channel Repeaters

The application of Digital On-Channel Repeaters (DOCRs) in DTV has been successfully demonstrated for two specific coverage conditions. These conditions include the filling in of coverage gaps caused by heavy shielding due to structures or terrain as well as the continuation of coverage over the horizon. If the receiving antenna of a DOCR has line-of-sight to the primary transmitter, it is likely that there will be more than 30 dB of received signal margin at the edge of the service area, thereby enabling provision of high quality signals to the area served by the DOCR. There are three basic types of practical DOCRs, each with distinct advantages and disadvantages. The types include: the RF amplifier or booster type, the IF conversion type, and the Equalization type. (The demodulation/re-modulation type briefly described earlier will not be discussed here as it is not considered practical for implementation in most SFNs.)

The RF amplifier or booster-type is the simplest. It processes the signal at RF and provides amplification and limited RF bandpass filtering.

Advantages: Lowest cost, simple, reliable due to minimal components required, low delay throughput, no STL is involved, and no second channel authorization is required.

Disadvantages: Susceptible to antenna feedback, limited output power, may not provide adequate adjacent channel rejection, no co-channel interference rejection, no multipath interference rejection, no noise reduction, no ability to adjust timing.

The IF conversion-type downconverts the RF signal to an IF signal that is bandpass filtered before upconversion and amplification.

Advantages: Low cost, simple, reliable due to minimal components required, no STL is involved, and no second channel authorization is required.

Disadvantages: Susceptible to antenna feedback, limited output power, may create long pre-echoes, may not provide adequate adjacent channel rejection, no co-channel interference rejection, no multipath interference rejection, no noise reduction, no ability to adjust timing.

The Equalization-type downconverts the RF signal to IF, demodulates, and applies equalization to extract baseband symbols. The process then is reversed, producing an amplified RF signal on the same frequency as at the repeater's input.

Advantages: Adjacent channel rejection, multipath interference rejection, noise reduction, may have short system time delay but longer than other DOCRs, no STL is involved, and no second channel authorization is required.

Disadvantages: More complex and higher cost than other DOCRs, susceptible to receiver input desensitization, limited output power but greater than other DOCRs, no ability to adjust timing.

5.2 Distributed Transmitters

The application of distributed transmitters is appropriate where it is desirable to maintain higher and more uniform signal levels throughout the coverage area. Distributed Transmission Systems also allow for the reduction of the transmitting antenna heights and the effective radiated power from the transmitter sites. While the DOCR is particularly effective wherever there is total

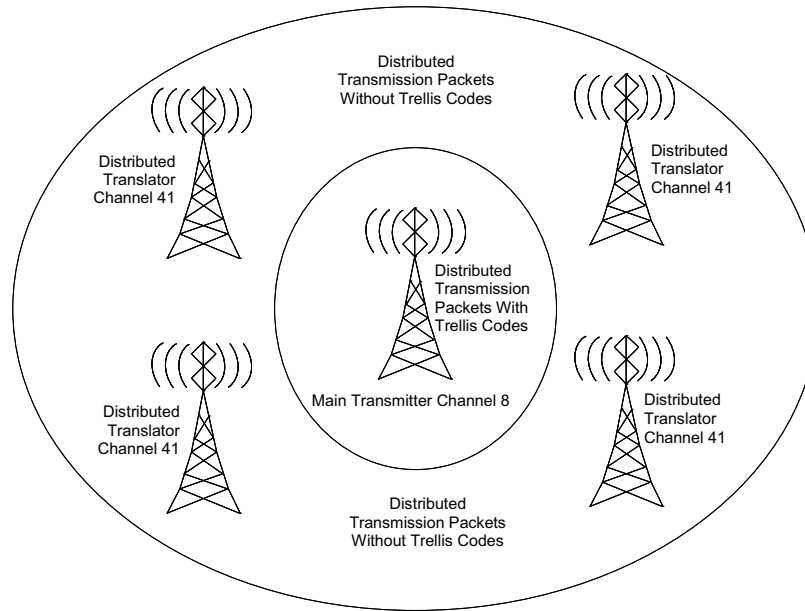


Figure 5.1 Single tier translator network using distributed transmission techniques.

blockage of the signal from the main transmitter, the Distributed Transmission System can operate under conditions of partial signal blockage.

Advantages: No second channel is required, output power is not limited, transmitter timing can be adjusted to optimize coverage.

Disadvantages: Most expensive approach, and an STL feed is required to each transmitter.

5.3 Distributed Translators

It is possible to implement a single-tier distributed translator network as illustrated in Figure 5.1. In this case the main transmitter, shown on Channel 8, acts as an “STL” to the distributed translators operating, in this example, on Channel 41. The main transmitter transmits the complete Distributed Transmission Packet (DTxP) to receivers at all of the distributed translators. The main transmitter would not be operating in a slave mode, since it does not require synchronization. The main transmitter simply passes the DTxP through, unmodified, with the trellis codes for synchronization intact. A single ring of translators can be added to extend the coverage area where necessary. The distributed translators, on the other hand, will remove the trellis codes from the DTxPs before retransmitting them.

ATSC A/110 [3] accommodates multiple hop distributed translator systems. It provides a technique that involves the insertion of multiple DTxPs — one for each tier of distributed translators. The Operation and Maintenance Packet (OMP) Type byte identifies the DTxPs for each tier. Each distributed translator removes the Trellis codes only from the DTxPs associated with its tier, and the DTxPs for other tiers are passed through unchanged.

Advantages: No STL feed is required to each translator, output power is not limited, transmitter timing can be adjusted to optimize coverage.

Disadvantages: A second channel authorization is required, and additional hardware is required at each translator to recover data not included in DTxPs when signals are relayed over the air.

5.4 Baseband Equalization Distributed Translators

If the signal from the main transmitter is clean and strong enough at the translator sites, it is possible to implement the distributed translator network shown in Figure 5.1 using baseband equalization distributed translators (EDTxR). Such translators down convert the signal to baseband (without decoding it) to apply equalization to compensate for channel distortions, and then up convert and retransmit on another RF channel. The pattern of the TCM symbols is not altered during this process and all translators emit the same signal without needing any Distributed Transmission Packet (DTxP). The output frequency of all the translators, however, should be precisely synchronized.

For the baseband equalization distributed translators to work, it is not necessary for the main transmitter to transmit Distributed Transmission Packets. However, if the output of the main transmitter contains such packets, they are passed through the translators unchanged and may be used in the next tier of distributed translators if necessary.

Advantages: STL not required, no DTxP required by the translators, output power not limited, possibility of transmission time adjustment for each translator for optimizing the coverage.

Disadvantages: A second RF channel assignment required, limited multi-tier operation due to accumulation of errors in the analog channels in the successive hops.

6 RECEIVER CONSIDERATIONS

The design goal for SFNs is to increase the number of locations where DTV service can be received. A kind of “self-generated multipath” is inherent in an SFN structure, however, and it can create some locations within the SFN service area where signal conditions created by the multiple transmitters actually make reception more difficult. The dominant cause of any increased reception difficulty is the higher level of multipath, especially a higher level of pre-echoes. Another difficult and likely condition that can be created is a “bobbing channel,” wherein the dominant signal path changes among two or more approximately equal echoes (which condition can also occur in systems using a single transmitter). High tolerance of large pre-echoes, multiple 0 dB echoes, and bobbing channels is normally associated with receiver design trade-offs in other important areas, such as S/N threshold.

6.1 Effects of Receiving Signals from Multiple Transmitters

Various reception conditions inherent in a Distributed Transmission Network can be problematic for reliable DTV reception. These conditions must be considered by the system designer in order to optimize the system. The most important such conditions are delay spread, channel fading effects, and Doppler shift effects.

6.1.1 Delay Spread

A consumer DTV receiver receiving signals from a Distributed Transmission Network will experience multiple signals arriving from the various transmitters in the network. The time of arrival of these signals at the receiver is dependent upon the relative distance between each transmitter and the receiver as well as the delay offset at each transmitter. Consequently, system

designers must be aware of these reception conditions and the impact on receivability. A particular concern is the possibility of significantly advanced echoes arriving at the receiver prior to the main signal. System designers must take into account the limited range of echo delays that can be handled by the DTV receiver.

In situations where STLs have a variable time delay (either due to use of satellites, propagation anomalies, or hardware characteristics), delay spread may be inadvertently increased. This may be remedied by using the GPS lock option (`stream_lock` flag = 0) discussed in Section 14.1.1.

6.1.2 Channel Fading Effects

A consumer DTV receiver receiving signals from a Distributed Transmission Network also may experience multiple signals from several transmitters that travel paths subject to fading. The effect of fading could create a situation in which signals from different transmitters can alternate as the dominant signal. This situation is most likely to occur when the received power at the receiver of each signal is nearly equal. System designers must be aware that this condition is difficult for DTV receivers and that reliable reception may not be possible.

6.1.3 Doppler Shift Effects

Doppler shift is a frequency offset caused by the movement of a transmitter and/or receiver or of nearby objects. Antenna sway, airplane flutter, moving automobiles, and mobile reception all cause Doppler shift effects. These situations can affect the signals from a single central transmitter or from SFN transmitters. Doppler shift effects result in dynamic multipath distortion that increases the DTV reception threshold or even can result in loss of reception.

In the synchronized multiple transmitter case, a frequency offset between transmitters can act like a Doppler shift from the perspective of a receiver. Therefore, it is crucial to maintain the frequency synchronization between SFN transmitters. The Synchronization Standard for Distributed Transmission A/110 requires a frequency stability for SFN transmitters of $\pm 0.5\text{Hz}$, which assures a very limited Doppler shift impact on consumer DTV receivers in a multiple transmitter environment.

6.2 Adjacent Channel Reception Issues

The implementation of a Distributed Transmission Network allows a broadcaster to expand its effective area of coverage. It is necessary, however, that system designers be cognizant of the potential for increased interference to receivers tuned to nearby or distant transmitters on adjacent channels. Since a distributed transmission network can place transmitters beyond the coverage limit of a single transmitter, power levels at a given receiver may be higher. The level of potential adjacent channel interference therefore can be increased.

6.2.1 Proximity of Transmitters to Receivers

In order to determine the potential impact of interference from adjacent channels, system designers must consider the expected performance of DTV receivers. The ATSC Recommended Practice on Receiver Performance Guidelines enumerates performance expectations for DTV receivers operating under various adjacent channel conditions. For first adjacent channel interference, the D/U planning factors for DTV interference into DTV are -26 and -28 dB in “Reconsideration of the Sixth Report and Order” (FCC 98-24). System designers must ensure that these planning factors are not exceeded.

Likewise, for taboo channels, system designers must ensure that the ratio of the desired and undesired signal levels does not fall below the performance limits of DTV receivers. ATSC A/74, “Recommended Practice: Receiver Performance Guidelines,” [2] also provides recommended performance levels for taboo channel interference. Table 6.1 tabulates taboo channel rejection thresholds for DTV interference into DTV. Table 6.2 tabulates taboo channel rejection thresholds for NTSC into DTV. System designers must ensure that these thresholds are not exceeded within the coverage areas either of a Distributed Transmission Network or of a transmitter on a taboo channel.

Table 6.1 Taboo Channel Rejection Thresholds for DTV Interference into DTV

Channel	Taboo Channel D/U Ratio (dB)		
	Weak Desired (-68 dBm)	Moderate Desired (-53 dBm)	Strong Desired (-28 dBm)
N+/-2	-44	-40	-20
N+/-3	-48	-40	-20
N+/-4	-52	-40	-20
N+/-5	-56	-42	-20
N+/-6 to N+/-13	-57	-45	-20
N+/-14 and 15	-50	-45	-20
Notes: All DTV values are average power.			

Table 6.2 Taboo Channel Rejection Thresholds for NTSC Interference into DTV

Channel	Taboo Channel D/U Ratio (dB)		
	Weak Desired (-68 dBm)	Moderate Desired (-53 dBm)	Strong Desired (-28 dBm)
N+/-2	-44	-40	-20
N+/-3	-48	-40	-20
N+/-4	-52	-40	-20
N+/-5	-56	-42	-20
N+/-6 to N+/-13	-57	-45	-20
N+/-14 and 15	-50	-45	-20
Notes: NTSC split 100% color bars with pluge bars should be used for video source. All NTSC values are peak power; all DTV values are average power.			

6.3 Receiver Characteristics

This section is intended to be cautionary, with the intent of increasing the awareness among SFN implementers of the possibly deleterious effects on receiver performance that might ensue. As such, it focuses on negative attributes and, for the most part, ignores the possible benefits of increased signal levels that SFNs can create.

6.3.1 Technical Characteristics

The elements of the receiver that are at risk from SFNs are mostly in the “front-end,” which includes the circuit elements from the input of the tuner to the output of the Forward Error Correction. Of particular concern are dynamic range, selectivity, synchronization, and, most especially, equalization.

Some general trade-offs include:

- Noise figure vs. adjacent channel overload capability
- Adjacent channel overload (tuner AGC) vs. impulse noise response
- IF selectivity vs. equalizer loading, which may also trade off against equalizer speed
- Large and distant pre-echo capability vs. S/N threshold

6.3.1.1 IF Selectivity

SFNs are likely to increase the level of adjacent channel signals to neighboring broadcasters at the outer reaches of the SFN coverage. This both affects the owner of the adjacent channel and puts pressure on receiver adjacent channel performance.

DTV receivers, like NTSC receivers, typically use superheterodyne principles in which the incoming RF signal is mixed down to an intermediate frequency (IF, typically centered at approximately 44 MHz), where most of the receiver's amplification and selectivity take place. In particular, adjacent channel frequencies are attenuated in the IF circuits. IF filters are made with surface-acoustic-wave (SAW) technology. SAW filters, unlike inductor-capacitor filters in early NTSC receiver designs, can be designed with linear phase response regardless of amplitude response. This controlled phase response is advantageous for digital reception.

The IF filter controls out-of-band attenuation and forms at least part of the in-band (desired channel) frequency shaping. Overall in-band channel response at the band edges must have a square-root-raised-cosine shape, per DTV specification. This overall shape is achieved by some combination of the SAW filter response, the response of digital filters usually residing in the demodulator integrated circuit (IC), and the equalizer. It is important that the SAW filter preserve the amplitude of the desired-channel band edges, certainly not rolling off more than the square-root-raised-cosine shape, in order to facilitate signal recovery. Because no SAW filter (or any other practical filter) can have an infinitely steep slope in the transition region between pass-band and stop-band, there are design trade-offs between a SAW that has maximum adjacent channel protection and one that preserves the in-band signal. The adjacent channel rejection of practical SAW filters in practical TV circuits is approximately 45–50 dB.

The ultimate stop-band attenuation of SAW filters is limited by a variety of practical effects. These include generation of acoustic waves in the bulk of the material (as distinct from the desired surface waves), reflections from the surface transducers, reflections from the edges of the piezoelectric substrate on which the surface transducers are placed, and stray electrical coupling that bypasses the acoustic device. State-of-the-art design techniques control and minimize these effects to improve the frequency response, but some of these mitigation techniques undesirably affect the previously stated independence of amplitude and phase responses. The result is the practical limitation on attenuation of adjacent frequencies.

Digital filtering is not necessarily a substitute for SAW filter selectivity because any undesired signals that get past the SAW filter must be processed by the analog-to-digital converter (A/D), and these extraneous signals tend to use up the dynamic range (number of bits) of the A/D.

In principle, the equalizer in DTV receivers can compensate for inaccurate SAW filter in-band shape and for tuner band-pass tilt, as well as do its primary job of canceling multipath., Reliance on the equalizer to make up for, e.g., band-edge attenuation by a SAW filter, however, incurs a noise penalty that could impair DTV signal recovery.

6.3.1.2 A/D Range and Resolution

The DTV signal can be converted from analog to digital form either at the output of the IF amplifier at ~ 44 MHz or after being mixed down to a frequency near baseband. Current state-of-the-art devices permit conversion at IF. For consumer devices at the sampling rates required by the bandwidth and equalization needs of DTV signals, 10 bits of A/D resolution is practical. Whatever this number is or might become, it sets an ultimate limit on the “digital noise floor” of the demodulator IC and is of particular consequence for equalization because it is one limitation on the amplitude of the multipath that can be cancelled with conventional equalizer designs.

The analog signal output of the IF amplifier must have amplitude scaled to match the input range of the A/D. Otherwise, a too-large signal will be “clipped,” resulting in serious distortion and ineffective signal recovery, or a too-small signal will use too few of the available 10 bits and suffer inadequate resolution for optimum processing in the demodulator filters, equalizer, and the like. The input to the A/D is “sized” by careful automatic gain control (AGC) in the tuner, the IF, and (perhaps) in a special scaling circuit associated with the A/D itself. These various AGC circuits are controlled by signal measurements within the demodulator, the IF, or the overflow bit of the A/D.

The practical dynamic range limitations of the A/D are important for considerations of adjacent channels, multipath, and other interference. In-band interference that cannot be eliminated by the tuner or IF (e.g., from DTV or NTSC co-channel, from adjacent channel “splatter”) adds to the desired DTV signal. Samples of the combined desired channel signal plus any interference must be converted “linearly” to numeric values by the A/D. If this conversion is not accurate, then any subsequent processing to mitigate the interference is doomed. Therefore, some portion of the A/D’s dynamic range is consumed by interferers, or equivalently, the number of bits devoted to the desired signal is reduced.

SFNs have the potential to increase the demands on the A/D undesirably in cases where the digital television receiver receives more strong signals than it can use advantageously. The main signal plus all added multipath signals must be converted accurately in order for the equalizer to work, and the total dynamic range must permit sufficient resolution of the samples even at frequencies severely attenuated by multipath.

6.3.1.3 Demodulator Issues

For the purposes of this section, “demodulation” encompasses all elements of the signal processing between the A/D and the FEC except for equalization (discussed separately below). Included functions are pilot tone acquisition and tracking, synchronization to pilot and data, AGC generation, rejection of NTSC co-channel interference, and recovery of the digital bit stream (and also equalization). There are various approaches — some proprietary — to each of these elements of the signal processing. The discussion herein will be confined to the underlying fundamental issues rather than specific processing details.

One of the basic design trade-offs is arithmetic resolution. This determines the “noise floor” of the digital processing. As described in the A/D discussion, the input is typically 10 bits resolution. The internal filters (e.g., for interference cancellation, band shaping, equalization, etc.) involve multipliers and adders and the resolution “bit width” internal to these functions may be higher than 10 bits. These design parameters are determined by overall simulation of the system, and they are scaled to maintain the threshold S/N ratio; i.e., the internal digital busses are made wide enough that increasing the bus width would bring uselessly diminished returns in the cost/performance trade-off. Although current IC technology permits high logic gate density and

designers use this to increase the arithmetic resolution of the processing, calculations to higher resolution than required can be very costly in circuit size and power consumed in some sub-systems.

The gate counts permitted on modern ICs are sufficiently high that performance trade-offs in the demodulation function are small. Compared with analog receivers, the digital demodulator assumes some of the interference rejection functions that were assigned to other areas (e.g., IF, tuner) in the past. Demodulation performance “limitations” are more a function of designer cleverness and knowledge of transmitted signal and interference characteristics than of cost or complexity constraints. Simulations of both the signal acquisition (i.e., the system) and of the IC (i.e., the detailed functioning of the exact gate-level design) are accurate and relatively rapid, and they enable effective design optimization.

The bottom line with respect to demodulator issues (other than those related to the adaptive equalizer) in an SFN environment is that there is very little impact expected on the performance of this portion of a DTV receiver from the use of SFNs as opposed to single transmitters.

6.3.1.4 Equalizer Characteristics

The specific function of equalization has been separated from the other demodulator functions because it is arguably the area of greatest concern for DTV receivers in an SFN environment and because the equalizer’s complexity and circuit size impose performance trade-offs. The DTV equalizer is, by far, the front-end circuit element with the highest logic gate count. Equalization is an area most DTV receiver manufacturers have emphasized in their development work to improve receiver performance. The additional performance required by SFNs, though, has not been a design goal for receivers presently marketed. Of particular concern is the nature of multipath in an SFN environment, in which large echoes, especially large and distant pre-echoes, can exist that exceed those found in the natural environment.

To avoid possible confusion, it must be understood that a simple statement of the number of taps in an equalizer is, at best, a partial description of the equalizer design and its performance. The time spacing of the taps (e.g., synchronous or “fractional” spacing) and whether the taps are “real” or “complex” are important elements of design and performance. In addition, Finite Impulse Response (FIR) filters, which are frequently employed to handle pre-echoes, generate a series of reduced-amplitude repeat “echoes” that themselves must be suppressed by the equalizer. Depending on the echo amplitude and time offset from the dominant signal, adequate suppression using an FIR design may require an equalizer considerably longer than the delay time of the actual transmission path echo. This receiver characteristic is one of the principal considerations that must be addressed in the design of an SFN.

Discussions of receiver performance and possible trade-offs are different for different aspects of multipath. Although all aspects are interrelated, separate discussions of amplitude, pre-echoes, post-echoes, and dynamic echoes illuminate the design considerations.

The quantization noise introduced during analog-to-digital conversion is a source of noise growth during the digital filtering associated with equalization. The equalization filtering seeks to combine differentially-delayed DTV signals so as to cancel component echoes, but, in the signal combining processes, the differentially-delayed quantization noise is likely to accumulate in the equalizer response. This adversely affects the SNR of the equalizer output supplied for data slicing. Additional quantization noise can be introduced within the equalizer filters in the digital multiplication of the differentially delayed DTV signals being weighted before their combination.

This type of noise growth can be controlled to some degree by ignoring the suppression of low-amplitude echoes.

In channels in which the principal signal is little stronger than a large number of echo components, such as conditions likely in SFNs, the problem of noise growth is compounded by the fact that FIR filtering in the equalizer attempts to cancel echoes with the filter input signal, which is not echo-free. Therefore, the echo spectrum cannot be reduced solely by FIR equalization filtering. In some designs, noise growth arising from repeated superposition of differentially delayed post-echo spectra is reduced with an Infinite Impulse Response (IIR) filter that suppresses post-echoes in the signals used for suppressing echoes in FIR equalization filtering.

In general, large-amplitude echoes can cause spectrum ripple that can seriously attenuate at least portions of the DTV signal spectrum. The equalizer attempts to restore the spectrum flatness, but noise amplification in the attenuated portions may reduce the restored signal's S/N ratio unacceptably. This may or may not be a problem with SFNs, depending on the ability of the SFN to deliver sufficiently high level signals to receivers. The situation is fundamental, however, and a solution depends on the SFN design, on a better antenna (improving the S/N), or on a steerable antenna (higher signal level, fewer or weaker multipath signals). Manufacturers have developed an EIA/CEA standard for an interface between a steerable antenna and a DTV receiver that provides for receiver management of the antenna automatically. Demodulator ICs are being offered that analyze the received signal and steer the antenna.

Cancellation of pre-echoes likewise faces some fundamental limitations. Typical state-of-the-art equalizers are time-domain filters. Stability of the circuits (a temporal issue related to the use of feedback in Infinite Impulse Response filters and imposed by the processing of signals that arrive *before* the desired signal) makes Finite Impulse Response (FIR) filters — i.e., feed-forward (as distinct from feedback) filters — the design of choice for handling pre-echoes. Unfortunately, feed-forward filters generate multiple, progressively-reducing-amplitude versions of the echo they are canceling. For large primary echoes, this increases the necessary length of the FIR filter in order to reduce all of the primary and secondary echoes to acceptable levels. In addition, FIR filters can substantially increase the noise level during the process of echo cancellation. Some of the noise is a result of finite arithmetic precision (as discussed above) and is mitigated by increasing the arithmetic bit precision in this local function. Some degree of noise growth is inevitable, however, with FIR filters. DTV receivers usually employ FIR filters to cancel pre-echoes as well as “close” (i.e., short-delay) post-echoes for a delay time equal to the designed pre-echo “advance” time. The noise growth issue with FIR filters imposes a fundamental trade-off between the pre-ghost time and amplitude that can be cancelled and the S/N damage imposed. Different manufacturers may declare different “best” points in this trade-off. Formal characterization of the DTx channel, in terms, for example, of differential delay between signals from multiple transmitters and within specific ranges of amplitude difference, would help — it would answer such questions as the practical range of pre-echoes.

Most descriptions of conventional (i.e., non-DTx) transmission channels show that post-ghosts can have longer time delays than pre-echoes are advanced. To handle post-echoes efficiently, Infinite Impulse Response (IIR) filters (i.e., feedback filters) are used. These filters do not generate repeat echoes, and they can be combined with a digital “slicing” or decision feedback operation to virtually eliminate noise growth. Processing of post-echoes can be improved by longer IIR filter structures in the equalizer, essentially without penalty other than circuit size and power consumption. While not the fundamental trade-offs associated with amplitude and pre-

echoes, the size and power penalties are not insignificant, given the number of gates in the equalizer.

“Natural” echoes are rarely static. It is expected that DTx echoes will also be variable, either from foliage, or structure sway, or passing traffic. Canceling echoes that move with time requires algorithms that process rapidly, that have some means of recognizing and measuring the echoes, and that compute updated equalizer parameters quickly. The transmitted DTV signal includes training waveforms that are useful for measuring echoes. Their utility for moving echoes is limited, however, by their low repetition rate. Virtually all state-of-the-art equalizers also track on the data itself, especially after initial acquisition using training waveforms. There are many algorithms for converging rapidly on the equalizer filter coefficients for canceling the received echoes, as found in text books and in the patent literature. Successful algorithms that can be implemented in consumer-level circuits are available in product.²

There are some trade-offs in Automatic Gain Control (AGC) operation between the performance that is optimum in moving multipath conditions and the performance desired in more stable signal conditions. Strong moving multipath can have a dynamic effect on received signal level, and controlled interaction that uses both the AGC system and the equalizer can sometimes be advantageous. A slower-acting AGC, however, can perform better in other channel conditions. The trade-offs are known to receiver designers, but different manufacturers may perceive marketing advantage in different “optimum” points. There are essentially no cost implications to these choices.

Frequency-domain equalization is also possible. In this case, a Discrete Fourier Transform (DFT) of the DTV signal is generated instead of the time-domain filter operations described above. Equalization is then performed in the frequency domain by normalizing the level of the signal across the channel. The control of noise growth offered by decision feedback equalization of post-ghosts, however, has led designers to favor the time-domain approach. Depending on filter length, time-domain processing may require fewer circuit operations.

6.3.1.5 Receiving Antenna Capabilities

Successful SFNs likely will depend on directional receiving antennas at least at some locations. The purpose of the antenna is to reduce selected multipath rather than to increase desired signal strength. As indicated above, receiver manufacturers have developed the EIA/CEA-909 antenna interface standard, which allows control and steering of compatible antennas by the demodulator in the receiver. The control is based on signal conditions measured or computed in the demodulator. Control algorithms attempt to steer the antenna pattern electronically to, e.g., minimize multipath as measured by equalizer activity. New demodulator designs include this capability. It is expected that new antennas will become available that support the standard.

6.3.2 Evolution of Characteristics

Receiver issues relating to multiple-transmitter-network environments differ with the sophistication of receiving devices. The level of sophistication has been, and is expected to

2. Additional information about receiver designs and the adaptive equalization ranges that may be encountered may be found in the ATSC Recommended Practice “Guide to the Use of the ATSC Digital Television Standard” (A/54A). Also see the ATSC Recommended Practice “Receiver Performance Guidelines” (A/74) and Technology Group Report “DTV Signal Reception and Processing Considerations” (T3-600).

continue, evolving over time. This evolution is addressed in three logical stages: legacy, current, and future receiver designs.

6.3.2.1 Legacy Receivers

Early receiver designs (generally, those designed and sold prior to 2004) are extremely vulnerable to aggressively designed SFNs, primarily because legacy receivers make trade-offs that do not expect the large and distant “unnatural” pre-echoes that SFNs can generate. By today’s understanding of the range of possible natural channel impairments, they tend to have very narrow equalization ranges, with very little capability for handling pre-echoes.

6.3.2.2 Current Receiver Designs

Current receiver designs (i.e., those created in 2004 and sold thereafter) offer some improvement over legacy product, primarily because of the steerable antenna interface. Equalizer performance also has improved, especially the handling of large, near-in echoes and distant post-echoes. Whether these improved features are adequate for SFNs remains to be seen.

6.3.2.3 Future Receiver Designs

Receiver design improvements that can be expected in the normal course of technology development are likely to aid reception of signals in SFNs. In addition, design trade-offs that emphasize SFN performance attributes are possible. Whether they will be favored will depend on the prevalence of receivers in locations served by SFNs versus receivers in “conventional” transmission environments.

7 APPLICATIONS OF DISTRIBUTED TRANSMISSION NETWORKS

At the most basic level, the application of a distributed transmission network can be divided into two categories:

- *Simple*, where a second transmitter is added to a main system.
- *Complex*, where multiple (three or more) transmitters are used in the system. Within the “complex” realm, systems can be further distinguished by “cell size.”

7.1 Simple — Adding a Second Transmitter

Basic operation of a Distributed Transmission System is illustrated by the use of two transmitters to create a Distributed Transmission Network, which has significant advantages when compared to traditional On Channel Repeaters and Translators.

On Channel Repeaters receive the signal that they retransmit over the air from a main transmitter. The most common design includes down-converting the signal to an intermediate frequency that is then amplified and converted back to the channel of the main transmitter. The result is a signal on the same channel with a small amount of time delay. Characteristics of On Channel Repeaters are:

- Power output is limited because of the potential for regeneration.
- Substantial isolation from transmitting to receiving antennas is required to avoid regeneration.
- Depending upon the design, received errors, channel impairments, and/or noise may be retransmitted.

- Time delay is added to the signal, which may adversely affect the performance of the consumer receiver equalizer.

Translators, on the other hand, require an additional channel, and spectrum may not be available for their use in many locations. Moreover, translators generally are not authorized for the higher power levels that are available to distributed transmitters.

A Distributed Transmission Network (DTxN) has many advantages when compared to On Channel Repeaters and Translators. In a DTxN, the digital bit stream is distributed over a separate path to each transmitter in the network. This eliminates both any errors that might have arisen using over-the-air distribution and the need for isolation between the transmitter and the over-the-air receiver. The result is that higher output power can be achieved. The transmitters also are synchronized, which allows control of the time delay between the two transmitters. Moreover, a second broadcast channel is not required, as it is when translators are used.

The major limitations of Distributed Transmission Networks are the network designs necessary to minimize mutual interference to consumer receivers (network internal interference) generated by the multiple transmitters and the limitations created by the operating ranges of consumer receiver adaptive equalizers. The impact of these limitations is in the distance by which the transmitters in a network can be separated from one another while still permitting the best possible improvement in reception.

The application of Distributed Transmission Networks can be illustrated by two examples related to terrain and coverage. In the first example, coverage is improved in terrain-shielded areas within a station's Noise Limited Contour (NLC). This use of a DTxN recognizes known coverage limitations by locating a distributed transmitter where it best can fill in an area receiving low signal levels because of terrain shielding. Since the Distributed Transmitters operate on the same channel, the spectrum allocations limitations on the use of translators do not preclude such service improvements.

The second example is the use of a second transmitter to maximize service in areas beyond the Noise Limited Contour of a single transmitter. In this case, the population to be served cannot receive the signal from the first transmitter, and the second transmitter is located so as to extend service to areas that otherwise could not be reached.

7.2 Complex — Multiple (3+) Distributed Transmitters

Complex Distributed Transmission Networks consist of multiple Distributed Transmitters and can use all of the concepts of the simple types discussed above. They have the ability to distribute power more uniformly throughout a total service area and to provide reception at any location from multiple directions. This can provide improved indoor reception. Signals arriving from several directions can be especially beneficial in structures with openings (windows) in only certain directions. Three concepts are closely related to the use of complex, multiple-transmitter applications: large cell systems, small cell systems, and micro-cell systems.

7.2.1 Large Cell Systems

Distributed Transmission Networks that use powerful transmitters on tall towers to cover large areas are called large cell systems. Such systems might have, for example, five cells each covering its own service area with a radius of 20–30 miles (30–50 km). The service areas of the individual cell transmitters would have large areas of overlap, and the network would require careful design to minimize the effects of delay spread on receivers in the overlap regions. Consideration of

transmitting antenna patterns, antenna elevations, transmitter power levels, and transmitter timing would be required. As DTV receiver adaptive equalizer designs improve over time, it can be expected that the constraints required on large cell designs and the difficulty of their implementation will be eased by the improved ability of receivers to handle the longer delay spreads that result from large cell networks.

7.2.2 Small Cell Systems

Distributed Transmission Networks that use lower power transmitters on shorter towers to cover smaller regions are called small cell systems. Such systems might have, for example, a dozen cells, each covering its own service area with a radius of 5–10 miles (8–16 km). The service areas of the individual cell transmitters would have relatively small areas of overlap, and the network would allow a simpler design while still minimizing the effects of delay spread on receivers in the overlap regions. Transmitting antenna patterns, antenna elevations, transmitter power levels, and transmitter timing would be designed to adjust the service areas to closely match the populations intended to be served. As DTV receiver adaptive equalizer designs improve over time, it can be expected that small cell designs can be further simplified due to the improved ability of receivers to handle the reception of multiple signals that result from small cell networks.

7.2.3 Micro-Cell Systems

Micro-cell systems use very low power transmitters to cover very small areas. They may be intermixed with either large cell or small cell networks to fill in coverage in places like city canyons, tunnels, or small valleys. Cities with tall buildings that emulate the effect of mountains and valleys with river gorges (e.g., New York and Chicago) can benefit from the use of microcells. Design of micro-cell systems will require the use of different design tools than typically used for broadcast applications — for example, those used for design of cellular telephone networks as opposed to the Longley-Rice methods used for longer range propagation modeling [4].

Ultimately, the application of Distributed Transmission Networks can use any or all of these concepts, integrated together in a master plan. This will allow evolution from a single, high power, tall tower system to a hybrid system based upon combinations of these concepts and other unique solutions created for each individual broadcast market.

8 DISTRIBUTED TRANSMISSION DESIGN CONSIDERATIONS

The design of a distributed transmission network involves a number of elements, not the least of which is the interference environment. This environment includes interference generated by the network itself, adjacent channel DTV stations, adjacent channel NTSC stations, and other sources. Multipath, of course, also comes into play and must be addressed during planning of the system.

8.1 Discussion of Interference Environment

As stated in Section 4.3.1, SFNs permit operation with less overall effective radiated power (ERP) and/or antenna height, as compared with the single central transmitter approach. This, in turn, leads to creating less interference to the service areas of neighboring transmitters that are not part of the network.

Another consequence of the use of SFNs is that different transmitters constituting the network may create “network-internal” interference to one another in certain regions of the station’s service area (see Section 4.3.2).

Contributions to the internal interference are considered to be only from those signals that carry the same information and are synchronized with one another. Otherwise, they fall into the category of external interference.

Various design methods can be applied to managing both external and system-internal interference. For example, selecting a specific network configuration that fits well with the topography of the service area and benefits from terrain shielding may minimize one or both types of interference. Antenna directivity may be applied to both cases, and delay adjustment of the transmitters may also be applied to mitigate system-internal interference. All such techniques should be considered when designing a single frequency network.

The following sections deal with the internal and external interference of SFNs in more detail, and introduce some ways to improve the operation of the network from the interference point of view.

8.1.1 Network Internal Interference

In the service area of a single, central transmitter, interference is created among the main signal and additional signals caused by the diffraction and/or reflection of the signal by stationary or moving objects in the vicinity of a receiving site. When these signals are combined, they form static or dynamic echoes with different amplitudes and delays, generally termed multipath distortion.

Because of the presence of multiple transmissions of the same signal in an SFN environment and the resulting system-internal interference, at any given receiving location, one can expect a more complicated situation than that of the single, central transmitter case.

8.1.1.1 Desired and Undesired Signals and Ratios

Interference between stations (external interference) is generally described in terms of a ratio between desired and undesired signal levels. In any particular case, the desired signal is the one being interfered with, and the undesired signal is the one causing interference to the desired signal. The ratio between the received signal levels of the desired and undesired signals at any given location will determine whether or not it is possible to receive the desired signal without unacceptable interference from the undesired signal.

In cases of external interference, the desired and undesired signals have completely different information content. Within the service area of either an SFN or a single, central transmitter, external interference is caused by neighboring transmitters operating on the same or different frequencies.

While the terms *desired* and *undesired signals* are widely used for external interference, they really do not apply well to network internal interference. For example, within a network, a signal may provide successful reception in one part of the service area but may contribute to an echo combination that is harmful and causes reception failure in another part. In this way, the signal may be desired at one point but undesired at another.

In dealing with the internal interference in an SFN, a practical approach would be to consider the multipath profiles within the SFN coverage area with respect to aspects such as multipath type, delay, and amplitude. Different multipath categories, each with a specific minimum required $C/(I+N)$, could be defined. Upon determining the multipath condition at each reception point, its corresponding minimum required $C/(I+N)$ could be found. This will be discussed in more detail in the following sections.

By considering multipath profiles, the network designer can also identify possible worst case scenarios. This, in turn, helps in selecting design parameters in the direction of improving those worst case situations. Consequently, one can expect that the SFN will have satisfactory performance for the majority of situations that may exist.

8.1.1.2 Different Signals in an SFN Environment

In order to simplify the design procedures for an SFN, it is helpful to define two types of signals, “natural echoes” and “SFN signals”.

“Natural echoes” are static and dynamic signals taking multiple paths from transmitter to receiver that are created by “diffraction” and “reflections” of signals from nearby (stationary and moving) objects.

“SFN signals” are the signals that a reception point receives directly from multiple transmitters in a single frequency network. They are usually considered in the design process for an SFN by using a propagation model to calculate their levels, from which their effects upon one another in a receiver can be predicted. All the secondary static and dynamic echoes resulting from SFN signals are considered to be natural echoes, which are not susceptible to prediction in the network design process.

Because various random and time varying factors are involved in causing natural echoes, it is not possible to accurately predict them at specific reception points. Consequently, a network designer does not have control over natural echoes; therefore, as one of the major challenges in DTV receiver design is to improve the receiver’s multipath performance, the impact of natural echoes in an SFN environment is addressed in other ATSC documents dealing with receiver considerations.³

The levels of SFN signals at reception points are usually much higher than natural echoes. Their contribution to the overall echo combination is sufficiently dominant that, in most cases, the impact of natural echoes can be ignored in the network design process. The network designer has control, however, over the amplitudes and delays of SFN signals and has the possibility of optimizing them.

8.1.1.3 Different Multipath Categories

Different transmitters in an SFN contribute to the creation of multipath distortion at some proportion of reception points. Since the key factors, from the receiver’s point of view, are the multipath delay and relative power level, it would be reasonable to categorize the multipath according to these two factors.

When multipath delay falls outside a receiver’s equalizer range, it acts like co-channel interference, i.e., additional noise in the channel, and is treated by the receiver as though it were not coherent with the other signals arriving at the receiver. For successful reception, the ratio of a dominant signal arriving at the receiver to the combined received signal levels of such multipath signals must be higher than the co-channel protection ratio characteristic of that receiver.

When the multipath delay is within a receiver’s equalizer range, the receiver should be able successfully to decode the signal. The existence of multipath, however, may require a higher C/N at the receiver’s input. The C/N margin is generally dependent on the ratio between the received signal levels of the dominant signal and of the multipath. The level of the multipath can be

3. See ATSC A/74 [2].

categorized into “weak”, “moderate”, and “strong” levels relative to the level of the dominant signal.

If the overall multipath level relative to the dominant signal is less than -10 dB, we may consider that the multipath distortion is weak. Levels of -10 to -6 dB correspond to moderate, and any level above -6 dB to strong multipath distortion.

DTV receivers should be able easily to deal with weak and moderate multipath conditions, but, when multipath distortions become strong, they may create a more challenging situation for the receiver.

From the network design and planning point of view, strong multipath conditions are expected to exist only within the overlapping coverage areas of the network transmitters. Their expected locations, however, can be predicted and mitigated, if necessary, by changing different design parameters such as transmitter power levels, tower location, spacing and height. Using a directional receiving antenna can also substantially reduce the multipath distortions. These issues will be discussed in more detail in the following sections.

For each category of multipath distortion explained above, we may consider a minimum $C/(I+N)$ ratio that is required for successful reception. In this ratio, “C” refers to the power sum of all the desired signals received at the reception point that fall within the receiver’s equalizer range. “(I+N)” refers to the sum of all undesired signals such as multipath distortions that fall outside the receiver’s equalizer range, all the external interference signals including co-channel and adjacent channel interference, plus the noise power in the channel from thermal and galactic sources. It should be understood that the “I” value under discussion includes only interference from the network itself and from neighboring television operations and does not consider unpredictable interference that occurs in the environment from other sources.

For the weak, moderate, and strong multipath distortions, the corresponding minimum required $C/(I+N)$ is taken to be 16 – 18 dB, 18 – 22 dB, and > 22 dB, respectively. Table 8.1 summarizes different multipath distortions that fall inside receiver’s equalizer range (both in delay and amplitude), and their corresponding $C/(I+N)$. For network design purposes, Table 8.1 summarizes the expected relationship between the amplitude ratios of signals received from different transmitters within a network and the required $C/(I+N)$ to permit reception by receivers whose performance follows the recommended guidelines in ATSC A/74 [2] Figure 4.3.

Table 7.1 Different Multipath Distortion Levels and their Corresponding $C/(I+N)$

Multipath Distortions Category	Weak, < -10 dB	Moderate, -10 to -6 dB	Strong, > -6 dB
Minimum Required $C/(I+N)$	$15.6 - 18$ dB	$18 - 22$ dB	> 22 dB

8.1.1.4 SFN-Signals Delay Spread

A reception point may receive a number of coherent SFN signals from different transmitters in the network. Among those signals, the strongest one may be considered as dominant and all others as pseudo-multipath signals.

The delay spread of the SFN signals at a reception point is defined as the arrival-time distribution of all the coherent signals. The signals that are below -20 dB (10 percent amplitude) with respect to the dominant signal may be considered negligible and not making any contribution to the delay spread.

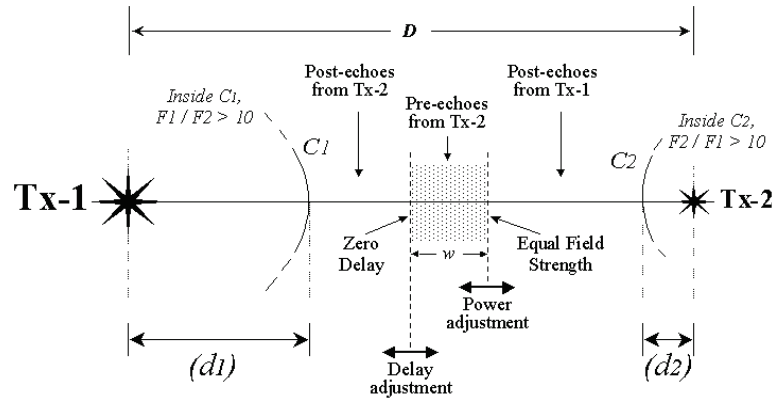


Figure 8.1 Internal interference zones between two transmitters in an SFN.

Taking the arrival time of the dominant signal at a reception point as the time reference, one can determine whether other signals are acting as pre- or post-echoes (i.e., falling pre- or post-cursor).

The time window of the delay spread extends from the arrival time of the first to that of the last SFN signal.

At any reception point, delay spread and its time reference (i.e., the relative location of the dominant signal — the cursor — within the delay spread) are determined by different factors, such as the network configuration and topology, as well as the emission timing of the different network transmitters.

Separation distances between the transmitters combined with relative delay adjustments made to the emission timing of the transmitters in the network determine the delay spread value at any location and affect its time reference. Larger separation distances usually result in wider delay spreads in parts of the service area of an SFN.

Transmitter power level, antenna height, and directivity can also affect the delay spread in the coverage area. By adjusting these parameters, one can change the relative location of a dominant signal in the delay spread, making other multipath signals appear as pre- or post-echoes.

8.1.1.5 Echo Situation in the Areas Between SFN Transmitters

When analyzing the echo situation in the areas between SFN transmitters, one should consider the fact that, in the vicinity of each transmitter, that transmitter's signal is strong enough to make the relative amplitude of other transmitters' signals negligible. Figure 8.1 illustrates this situation more clearly for a simplified case of an SFN consisting of only two transmitters.

In Figure 8.1, $F1$ and $F2$ denote the field strengths of the signals from transmitters Tx-1 and Tx-2. Tx-1 is assumed to be more powerful than Tx-2. $C1$ and $C2$ are the contours on which the ratios between the field strength of the nearby and distant transmitters are 20 dB. D is the separation distance between the transmitters, and $d1$ and $d2$ are the distances from the respective transmitters to their associated contours along the line between the two transmitters. Inside a contour, the relative field strength of the distant transmitter's signal is less than -20 dB (field strength ratio of 1/10) relative to the nearby transmitter's signal. Under these conditions, within the contour, the signal from the distant transmitter is not strong enough to make any contribution to delay spread or to create any network internal interference.

Distances d_1 and d_2 depend on the effective radiated powers (ERPs) and antenna heights above average terrain (HAAT), as well as the separation distance between the transmitters D . Longer separation distances result in longer d_1 and d_2 values.

Also shown in Figure 8.1 are two lines, one indicating the locations of zero delay between the arrival times of the signals received from the two transmitters and the other one corresponding to the locations of equal field strength of the two signals (i.e., zero dB echo).

If no additional delay is applied to the emissions of either of the transmitters, then the zero delay line will fall in the middle, equidistant from the two transmitters. By delaying or advancing the emissions of one of the transmitters relative to the other, however, the line can be moved toward one or the other of the transmitters and away from the other one.

Depending on the propagation model used to predict the field strength and whether or not terrain irregularities are taken into account, the shape of the equal-field-strength curve may vary. For simplicity, it is shown as a straight line in the Figure 8.1. By changing the ERP or the HAAT of either transmitter, this line also can be moved back and forth between them.

Concerning the post-echo situation, one can see that on the left side of the zero delay line, the Tx-1 signal is stronger and dominant, and the Tx-2 signal acts as a post-echo because it reaches the area later than the Tx-1 signal. By the same analysis, on the right side of the equal-field-strength curve, the Tx-1 signal acts as a post-echo. In Figure 8.1, these two areas are designated as the areas of post-echo from transmitters Tx-1 and Tx-2.

Upon moving from the zero delay line toward either of curves C_1 or C_2 , the delay of the echo signal will increase. The maximum delay value of the post-echo signal is the time corresponding to twice the distance from the zero-delay line to either C_1 or C_2 , whichever is longer. The distance must be doubled because, when moving toward one transmitter, one encounters the double effect of getting closer to that transmitter and, at the same time, getting farther from the other one. For the case shown in Figure 8.1, because no delay is applied to either of the transmitters, the zero-delay line is at the center of span D , and has a longer distance to curve C_2 . Under these conditions, the maximum delay of the post-echo signal is the time corresponding to $2(D/2 - d_2) = D - 2d_2$, or, in microseconds, $(D - 2d_2) / 300$, where D and d_2 both are in meters. This is the width of the post-echo portion of the time window of the delay spread.

If a signal delay of $(d_2 - d_1) / 300 \mu\text{s}$ (where d_1 and d_2 are in meters) were applied to Tx-2, then the zero delay line would be pushed to the center point between C_1 and C_2 . Under this circumstance, the width of the post-echo portion of the time window would be reduced to the time corresponding to the distance between C_1 and C_2 ; i.e., $(D - d_1 - d_2) / 300$. This is the minimum value that the post-echo portion of the time window of the delay spread can have.

Concerning the pre-echo situation, the area between the zero-delay line and the equal-field-strength curve is the only area in which the dominant signal (from Tx-1) arrives later than the echo signal (from Tx-2), resulting in the formation of pre-echo (signal arriving in advance of the dominant signal). This area is shaded and designated as the pre-echo area (from Tx-2) in Figure 8.1. The maximum advance of the pre-echo signal occurs immediately adjacent to the equal-field-strength curve, with a time value corresponding to twice the width of the pre-echo area w . In microseconds, it is $2w / 300$, where w is in meters. This is the width of the pre-echo portion of the time window of the delay spread.

By adjusting the emission delay, power, and antenna height of the transmitters, the relative positions of the zero-delay line and the equal-field-strength curve — and consequently the

position and width of the pre-echo area — can be adjusted. For example, it may be pushed to unpopulated areas, or its width even may be reduced to zero.

8.1.1.6 Separation Distance Between SFN Transmitters

When designing an SFN, one should bear in mind that various design parameters such as separation distance, ERP and HAAT of the transmitters, maximum delay of pre- and post-echoes, and the like, are all related to each other. Any change made to one of these parameters causes other parameters also to change.

For example, considering the simplified SFN that is shown in Figure 8.1, d_1 and d_2 would be enlarged if the separation distance D could be increased. If the relative delay of the signals should not be more than a certain limit in the areas between curves C_1 and C_2 , however, then the separation distance between the transmitters could not be increased beyond a certain value.

With no additional delay applied to the transmitters, the relationship between the maximum delay of post-echoes τ and the separation distance D was already found to be $\tau = 2(D/2 - d_2) / 300$ or $D = 300\tau + 2d_2$. In this relationship, distances are in meters, and time is in μs . If τ is the maximum permissible delay, then D obtained from the above relationship would be the maximum separation distance.

By applying sufficient delay to Tx-2 to push the zero-delay line to the center point between C_1 and C_2 , the separation distance could be increased up to $300\tau + (d'_1 + d'_2)$, where d'_1 and d'_2 are the new distances that result from the increased separation distance. When applying such a delay, other parameters also must be adjusted to avoid unacceptable pre-echoes.

If the maximum permissible delay can be increased or if the areas within which such delay is exceeded are not populated, then the separation distance may be further increased.

Another factor affecting the separation distance is the threshold field strength ratio of 20 dB considered for curves C_1 and C_2 in Figure 8.1. Under certain conditions, this ratio can be reduced to enable an increased separation distance. For example, if the number of echoes is not high, if terrain irregularities may provide some shielding, or if a low gain directional receiving antenna is assumed to be used in some areas, then the above-mentioned field strength ratio may be reduced to some extent.

8.1.1.7 DOCR vs. Other Sources of Transmission in an SFN

As its input signal, a Digital On-Channel Repeater (DOCR) receives a signal over the air from a main or distributed transmitter. (Throughout this section, a main transmitter will be discussed, but the same concepts apply to either type of signal source for the DOCR.) The signal is processed within the repeater and retransmitted toward a target area.

With respect to the output signal of the main transmitter, the output signal of a DOCR has a delay that is equal to the amount of time required by the signal to travel from the main transmitter to the repeater plus the internal delay of the repeater. This delay has a specific effect on the delay spread in the coverage area of the repeater.

Relative to other sources of transmission in an SFN, DOCRs may have more pre-echoes in their coverage areas. For example, if a main transmitter's signal could also be received in the coverage area of a DOCR, it would always be received sooner than the DOCR's signal. Under these circumstances, in the vicinity of the DOCR, where its signal is dominant, the main transmitter's signal would act as a pre-echo.

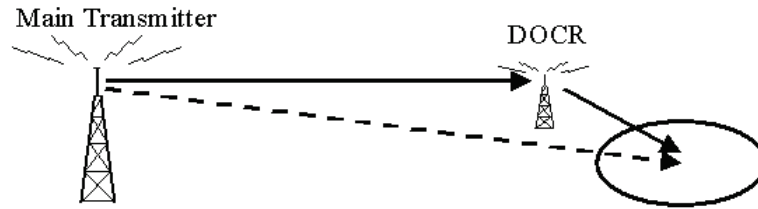


Figure 8.2a Main transmitter and DOCR signals in the repeater's coverage area.

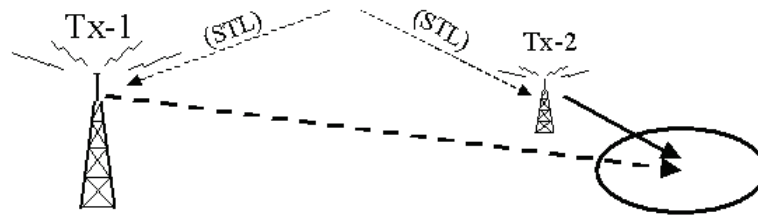


Figure 8.2b SFN transmitters' (Tx-1 and Tx-2) signals in Tx-2 coverage area.

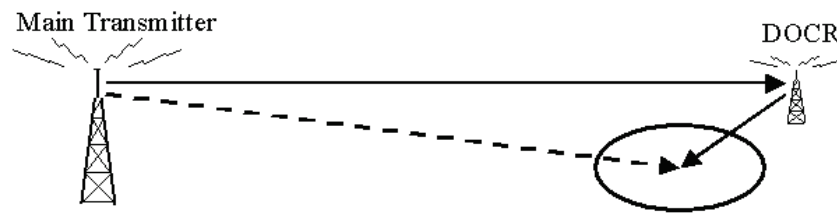


Figure 8.2c Higher delay due to path difference between the main transmitter and DOCR signals in the repeater's coverage area.

Figure 8.2a illustrates the DOCR situation. The solid line shows the path of the signal through a DOCR, and the dotted line indicates the direct path from the associated main transmitter to a reception point in the coverage area of the repeater. In addition to traveling a longer distance, the repeated signal also encounters the DOCR internal delay when passing through it.

Figure 8.2b shows a similar configuration, but instead of a main transmitter and a DOCR, there are two SFN transmitters (Tx-1 and Tx-2) fed by two studio-to-transmitter links (STLs). In contrast with the DOCR case of Figure 8.2a, if the two transmitters are transmitting with no delay with respect to one another, the Tx-1 signal will reach the receiving point later than that of Tx-2 and will act as a post-echo when the Tx-2 signal is dominant.

An important objective in designing a system using a DOCR should be to minimize the difference in path lengths between the direct signal from a main transmitter and the signal coming through the repeater. This objective can best be achieved by locating the DOCR between the main transmitter and the DOCR's intended coverage area, as in the configuration shown in Figure 8.2a. For other configurations, such as shown in Figure 8.2c, in which the DOCR coverage area lies somewhere between the main transmitter and the repeater, the difference in path lengths traversed by the two signals and, consequently, the delay spread between them in the repeater's coverage area would be much longer. The longer path delay, combined with the internal delay of the

repeater, may result in the main transmitter's signal being received as an earlier pre-echo in the repeater's coverage area.

Taking the various considerations into account, it is best to use a DOCR to cover shadowed areas that do not receive a strong signal from any other source in an SFN. Overlapping of coverage areas between a DOCR and other SFN transmitters should be avoided to the extent possible. If, it is not possible to avoid such overlapping areas in which a repeater's signal is dominant, then using a repeater with low internal delays in a network configuration having lower differential delay (such as shown in Figure 8.2a) would result in shorter pre-echoes.

8.1.1.8 Signal Directivity

As described in previous sections, managing the internal interference in an SFN may require control of the effective radiated power (ERP) from the various transmitters toward the areas in which those transmitters have overlapping coverage. When factors such as terrain and the location of population centers are considered, different ERP values may be required in different directions from each transmitter. This sort of signal directivity can be achieved through shaping of the transmitting antenna radiation pattern, taking into account the relative height of the antenna with respect to terrain in each direction.

It should be noted that network internal interference is not the only factor that must be considered when evaluating ERP adjustments. Other factors include SFN coverage requirements and the external interference that may be created both inside and outside the network service area.

As previously described regarding the application of DOCRs, in the areas between the main transmitter and the repeater, the longer delays due to path differences between the signals from the two sources may cause very long pre- or post-echoes. To mitigate this situation, the DOCR can be situated so that it can serve the intended coverage area by transmitting away from the main transmitter. In these circumstances, radiation toward the main transmitter should be limited, and this limitation can be achieved by using a directional transmitting antenna for the DOCR.

In addition to directional transmitting antennas, the use of receiving antenna directivity also may have a significant impact on successful reception at certain locations. Consequently, the assumption of use of directional receiving antennas is an important consideration among SFN design parameters. In practice, adjustment of the direction of a receiving antenna that has a few dB of gain (or a notch), may reduce the relative amplitude of an echo from strong to moderate at the output of the antenna.

Strong echo conditions caused by the signals from multiple transmitters are likely only in portions of the service area of an SFN, and network design parameters should be optimized to provide successful reception in those areas. By assuming use of a directional receiving antenna, the toughest receiving conditions that might be predicted to exist in these areas can be mitigated to some extent, resulting in a consequent relaxation of limitations on design parameters that might otherwise be necessary.

In indoor reception, there may exist in one location SFN signals from several transmitters, with potentially large standing wave effects. A directional receiving antenna may be particularly useful in such situations when compared to an omnidirectional antenna by helping the receiver to choose one of the signals as the dominant signal.

It should be noted that, for receiving antennas, the important parameters are directivity and steerability. The antenna should be able to attenuate (by even a few dB) an echo signal that is coming from a direction different from that of the dominant signal. An antenna having some nulls

may be of more use in such situations than one with very high gains or front-to-back ratios. Such requirements may make the design of the receiving antenna somewhat less critical.

8.1.1.9 Terrain Shielding

In an SFN, overlapping of the coverage areas of different network transmitters is an expected outcome. Such overlapping coverage areas may be intentionally planned, or unplanned and unintentional.

There may be situations in which portions of the coverage areas of SFN transmitters are deliberately planned to overlap. Multiple network transmitters usually cover such areas of overlap from different directions. Downtown canyons are typical candidates for locating intentional overlapping coverage areas. If covered by a single transmitter, they would suffer from too many shadows formed behind the high-rise buildings.

In such downtown canyon situations, reception on the sides of the high-rise buildings facing away from a single transmitter usually involve signals reflected from other nearby buildings. If the direct signal can reach the opposite side by passing through a building, it generally will encounter a high penetration loss and thus most likely act as a pre-echo. Depending on the situation, the reflected signals may be numerous and even time varying, and can create a complex echo combination that prevents the receiver from achieving successful reception.

If another SFN transmitter is positioned to illuminate the shadowed side of the high-rise buildings, it can provide a more powerful dominant signal for the receivers on that side, thereby giving them a better chance of dealing with echoes and achieving successful reception.

When designing an SFN, planning parameters are generally selected to provide favorable reception conditions in the intentional overlapping coverage areas. Such selections, however, may have the potential of creating unintentional overlapping signal areas in other locations.

If the situations in unintentional overlapping coverage areas are unfavorable, e.g., they result in harmful echo combinations, then they should either be avoided or be prevented from falling within populated regions.

Careful selection of network design parameters can mitigate unfavorable conditions in unintentional overlapping coverage areas or change their locations. It may be necessary to trade off improvements in such overlapping coverage areas so as not to adversely affect other portions of the service area. One very efficient way in which to avoid unintentional overlapping areas is to make use of terrain irregularities as shields between network transmitters.

Terrain irregularities should be taken into consideration from the very beginning of the SFN design process. They can help in choosing better sites for transmitters and in making the choice between types of network transmitters. For example, if an area were shielded by terrain from other transmitters in the service area of an SFN, then it would be an ideal candidate for using a DOCR. The shielding effects of terrain are an additional parameter that can be used, along with others such as signal directivity, ERP, delay adjustments, and the like, to achieve a variety of design goals in SFNs.

8.1.2 Impacts on Coverage of Other Stations

Managing interference in designing an SFN is a two-way process and should be considered bi-directionally. In one direction, the interference may be received within the service area of an SFN from neighboring stations that are not part of the network and are operating on the same or adjacent channels. These stations may be in the same market or in neighboring markets. In the

opposite direction, interference may be created by the SFN transmitter(s) in the coverage areas of those neighboring stations. The latter, referred to as SFN external interference, will be considered throughout this section.

One possible approach for studying the impact of an SFN on the coverage of other stations would be to compare the SFN, from the external-interference point of view, with a single-central-transmitter. Based on such a comparison, the differences that may exist between the two situations may be highlighted and their impacts on interference calculations determined and accounted for.

In this way, it should be possible to modify the existing methods in use for single central transmitter cases and to use them for external interference considerations with respect to an SFN. Existing methods and their derivation criteria may be different; however, for different countries and administrations and may require different modification procedures.

8.1.2.1 Within Market Stations

Stations within the same market are those using either single or multiple transmitters whose service areas overlap one another at least partially.

When the impact of an SFN on other stations is considered, there should be a distinction made between the transmitters that are within the same market but operating on separate frequencies, and the transmitters that are part of the SFN. The mutual effects of SFN transmitters on each other are considered in Section 8.1.1.1. In this section, the impacts of an SFN on other “within market” stations will be studied.

8.1.2.1.1 Adjacent Channel DTV

Development of a DTV allotment plan is based on avoiding first adjacent channel assignments within the same market to the extent possible. Under the circumstances that such assignments are unavoidable, usage of the same site and co-location of the adjacent channel DTV transmitters generally forms the basis for planning single central transmitter cases.

In the single transmitter case, if the ERPs of the two co-located, adjacent channel DTV transmitters are not equal, then the more powerful transmitter may create harmful interference in the coverage area of the weaker one. In order to avoid such interference, there generally is a limit on the maximum difference between the ERPs of the two transmitters.

Taking the ERP of the lower power transmitter as a reference, the maximum permissible ERP of the more powerful transmitter can be found based on the required D/U ratio. For example, for upper and lower first adjacent DTV into DTV, D/U should be at least -26 and -28 dB, respectively. This means that the interfering, undesired, upper adjacent DTV signal level can be 26 dB stronger than the desired DTV signal level before it can cause harmful interference. The ERP that can produce such an interfering signal level at the edge of the lower power DTV coverage area is the maximum limit for the higher power DTV transmitter. It should be noted, however, that any calculation for determining the desired and undesired signal levels should be based on the location and time availability corresponding to coverage and interference, respectively, and also should take into account antenna pattern differences between the stations.

When adjacent channel DTV transmitters are co-located, their signals pass through the same paths to reach any reception site. As their frequencies are close to each other, their respective channel characteristics are likely to exhibit high correlation between them.

If co-location is not possible, a separation distance of up to a few kilometers (e.g., 5 km in the U.S.) may be allowed between the locations of the two adjacent channel DTV stations. Under this

circumstance, the limit on the maximum difference between the ERPs of the two DTV transmitters is reduced. This is because, under the new conditions, the stronger station becomes closer to parts of the edge of the weaker station's coverage area and, as compared to the co-location case, can create a stronger interfering signal at those locations. Reduction in the maximum permissible ERP of the stronger DTV station compensates for the additional strength of its signal at such locations.

If an SFN is to be used in a single market by one of two (or more) adjacent channel stations, the best approach is that both (all) the adjacent channel stations operate as SFNs and co-locate their transmitters. This allows avoidance of the problem of D/U ratios exceeding the limits of adjacent channel performance that can be expected from consumer receivers and that generally are built into the rules controlling the service.

If only one of two adjacent channel DTV assignments were configured as an SFN, one can expect different SFN transmitters with different ERPs at different locations in the coverage area of the single central transmitter. When designing an SFN under such circumstances, the following considerations must be taken into account.

- Based on the same rationales pertaining to the situation in which both adjacent channel assignments use single central transmitters, co-location of one of the SFN transmitters with the single central transmitter is highly recommended. All the rules and procedures that are relevant to co-located single central transmitters are also applicable to such an SFN transmitter. Other SFN transmitters sited at other locations, however, must be considered very carefully.
- By the same reasoning stated above for a pair of single central adjacent channel transmitters, as the separation distance between an SFN transmitter and the single central station increases, maximum permissible ERP of the SFN transmitter decreases. Depending on the separation distance, such a maximum permissible ERP may become less than the ERP required by the SFN transmitter for providing satisfactory service in its intended coverage area.
- By limiting the radiation of the SFN transmitter in the direction of the edges of the coverage area of the single central transmitter, its maximum permissible ERP in the direction toward the single central transmitter may be increased. In addition to reducing the ERP in the direction of the edges of coverage of the single transmitter, exploiting the shielding effects of terrain irregularities and natural obstacles with respect to the SFN transmitter's signals can also limit the interference effects of the SFN transmitter in the coverage area of the single central transmitter. This, in turn, can result in a further increase in the maximum permissible ERP of the SFN transmitter.
- Depending on the ERP of the SFN transmitter, when its separation distance from the single central transmitter is increased beyond a certain limit, another interference spot begins to form in the vicinity of the SFN transmitter. In this area, the SFN signal is strong enough to prevent reception of the single central transmitter's signal. This is because, as one gets farther from the single central station, its desired signal becomes weaker and more susceptible to interference from the SFN transmitter's signal; in the vicinity of the SFN transmitter, the required D/U ratio can no longer be obtained. If the SFN transmitter site is chosen to be at a location in an unpopulated area, formation of such an interference spot can be tolerated to the extent that it does not fall over the adjacent populated areas. Interference hot spots of the sort described surrounding SFN transmitters may be

controlled using antenna pattern characteristics that greatly reduce the signal level from the SFN in the area under its antenna.

- Depending on the situation, either the interference at the edge of the coverage area of the single central transmitter, or the formation and extent of an interference spot around an SFN transmitter may become dominant and cause a reduced limit on the maximum permissible ERP of the SFN transmitter.
- Limitation of the maximum ERP of distant SFN transmitters may be the most serious constraint on SFN implementation imposed by a single central adjacent channel transmitter and should be carefully considered in the process of designing an SFN.
- There may be cases in which the ERP calculated as necessary to provide acceptable service in the intended coverage area of an SFN transmitter substantially exceeds what is permissible from the standpoint of interference that it would cause. In such a case, one possible solution is to split the intended coverage area of the SFN transmitter into a number of smaller cells, each of them covered by a lower power SFN transmitter.
- In order to resolve the mutual interference problems between an SFN and a single central adjacent channel transmitter within the same market, a compromise may be required between the distance separating an SFN transmitter from the single central adjacent channel transmitter and the SFN transmitter's coverage area, as determined by its ERP. Studies and coordination procedures may be carried out to determine whether it may be possible to tolerate affecting to some extent the coverage of the single central station at the edge of its protected contour or in the vicinity of the SFN transmitter.
- In performing any calculations of interference from transmitters operating in an SFN, the cumulative effect of the signals originating from the several transmitters in the network must be taken into account.

8.1.2.1.2 Adjacent Channel NTSC

In the transition period from analogue to digital TV broadcasting, NTSC and DTV adjacent channel allocations may be used in the interim plan. For the single central transmitter case, the interim plan normally is based on using existing sites, co-locating the NTSC and DTV transmitters, and replicating the coverage of the existing NTSC station. Based on these assumptions, the required ERP for the DTV transmission can be calculated.

All the considerations relevant to adjacent DTV/ DTV channel assignments discussed in the previous section are also applicable to adjacent NTSC/ DTV channel assignments. Based upon the same type of analysis given above for adjacent channel DTV transmitters, in order to avoid harmful interference to an existing NTSC assignment, the ERP of a co-located adjacent channel DTV transmitter must be limited so as to maintain the required D/U ratio. Furthermore, when co-location of adjacent channel NTSC and DTV transmitters is not possible, a maximum separation distance between the two stations is allowed, and consequently, the limit on the maximum permissible ERP of the DTV transmitter with respect to the NTSC transmitter will be reduced accordingly. The major distinction between the two cases of adjacent DTV/DTV and adjacent NTSC/DTV assignments, however, is the higher value of protection ratio required by NTSC signals with respect to adjacent channel DTV signals.

For example, the minimum D/U ratio permitted for upper first adjacent channel DTV into NTSC is -12 dB^4 . When compared with the permitted D/U of -26 dB for upper adjacent DTV into DTV, one can conclude that NTSC puts more restrictions than does DTV on the maximum

permissible ERP of an upper adjacent channel DTV transmitter. If a DTV transmitter is operating on the upper adjacent channel of an NTSC assignment, its maximum permissible ERP will be 14 dB lower than that of the same transmitter operating under the same conditions on the upper adjacent channel of another DTV transmitter having the same ERP as the NTSC transmitter.

When a DTV assignment adjacent to an NTSC station is implemented with an SFN, larger separation distances between the NTSC and the SFN transmitters can be expected than if the NTSC's neighbor were a single central DTV station, and such a tight constraint imposed by NTSC may cause a significantly lower limit on the maximum permissible ERP of distant SFN transmitters.

Considering the D/U ratios of – 49 and – 48 dB required by DTV against interfering upper and lower first adjacent channel NTSC signals,⁵ respectively, the interference effect of NTSC on DTV is insignificant.

Another important issue to consider in NTSC/SFN first adjacent channel assignments is the possible degradation of the NTSC picture quality inside the NTSC coverage area. Such reductions may be within technically acceptable limits but, at the same time, may cause undesired consequences. The minimum required D/U ratios discussed above for protection of NTSC signals are all based on reception of NTSC signals with CCIR Grade 3 picture quality. In other words, if the ratio of the desired NTSC to undesired DTV signals at a reception point is equal to the minimum required D/U ratio, one can expect a Grade 3 NTSC picture quality at that point.

When adjacent channel DTV signals are from a single central transmitter co-located with an NTSC transmitter, one should be concerned about the edges of the NTSC coverage area as the areas that may first be affected by DTV interference. Protecting NTSC signals in these areas would result in lower interference effects from DTV in the areas that are closer to the NTSC transmitter location. At the edges of the NTSC coverage area, however, the NTSC signal strength is already close to its lower limit, and the picture quality is more likely close to Grade 3.

When a DTV adjacent channel assignment is implemented with an SFN, one can expect several SFN transmitters to be sited at different locations in the NTSC coverage area, even inside its grade A or city grade contours. Installation of SFN transmitters in such areas may affect the received NTSC quality grade, especially in the vicinity of an SFN transmitter. Even though this may be technically tolerable, it may cause severe reception degradation for those who previously had been able to receive the NTSC signal with a very high picture quality. This may not be acceptable to those viewers or the NTSC broadcaster.

Taking all of the preceding considerations into account, implementation of an SFN in the presence of an NTSC first adjacent channel is a very challenging task and requires very special care. Under certain circumstances, it may not be possible to operate some SFN transmitters at their nominal power levels during the NTSC / DTV transition period.

8.1.2.2 Neighboring Market Stations

Other stations that operate outside the service area of a given station and do not share an overlapping coverage area with that station are considered to be neighboring market stations. (Note that, even if two stations do share some overlapping coverage area, the areas they serve without overlap can be treated in the same way as neighboring market stations.) If the separation

4. See Appendix A of the 6th Report And Order in MM Docket No. 87-268 FCC 97-115.

5. Id.

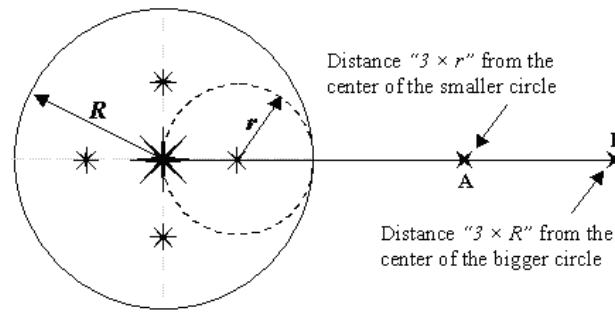


Figure 8.3 Interference limits of a single central transmitter and of an SFN.

distance of a neighboring market station is less than some limit, it may create undesired effects in the service area of the station under consideration.

When the impact of SFN transmitters on other stations is considered, a distinction should be made between the effect on transmitters that are in a neighboring market and operating on the same frequency as the SFN and on other transmitters that are part of the same SFN. The mutual effects of SFN transmitters in a network on each other were considered in Section 8.1.1. In this section, the impacts of an SFN on neighboring market stations operating on the same or different frequencies will be studied.

8.1.2.2.1 Co-Channel

If a single central transmitter were replaced by a number of SFN transmitters, the transmitted power would be distributed across the coverage area, and, in this way, interference that is created for neighboring market stations could be reduced. This result can be demonstrated with some simple calculations and an example.

Based on calculations for location and time availability using the FCC F(50,10) curves for interference and the F(50,90) curves for coverage, it may be shown that co-channel interference from a digital UHF transmitter will extend on the order of three times the distance over which it can provide coverage (see Section 5.1).

Figure 8.3 shows a very simplified example of four SFN transmitters together providing approximately the same coverage as that of a single central transmitter. In this figure, the bigger star represents a single central transmitter at the center of its circular coverage area having a radius R , and the four smaller stars represent four SFN transmitters each having a coverage area of radius $r = R/2$. Point "A" is at a distance of $3 \times r$ from the closest SFN transmitter and point "B" is at a distance of $3 \times R$ from the single central transmitter. Replacing r with $R/2$, the distance of point "A" is found to be $2 \times R$ from the single central transmitter.

Neglecting the cumulative effects of the more distant SFN transmitters, these calculations result in point "A," at the distance of $2R$ from the center of the bigger circle, being the limit of the interference zone of the SFN in the specific direction shown. As the limit of the interference zone of the single central transmitter is point "B," at a distance of $3R$ from the transmitter, one can conclude that, in this particular case, using an SFN results in reducing the radius of the interference zone to a distance of approximately $2/3$ that of the single central transmitter.

In this simplified example, if an SFN transmitter were moved toward the edge of the coverage area and its ERP reduced in that direction, then its interference to a station in the neighboring market would be reduced further. It is important to note that replacing a single central transmitter with an SFN also has the effect of making reception in most parts of the service area more robust against interference from neighboring stations. Distribution of power among different SFN transmitters makes the SFN signal strength more uniform across the service area and more tolerant to stronger unwanted signals from neighboring stations.

The example above shows that, when an SFN is replacing a single central transmitter, its design can have the effect of reducing interference to neighboring market stations. It should be noted, however, that the DTV allotment plan already in existence is based mainly on single central transmitters that are replicating the coverage of existing NTSC stations. Under these circumstances, design and implementation of an SFN is actually a replacement for an already planned single central DTV transmitter and may have to respect its protected contour and coverage area. For a planned single central transmitter, adequate separation distances from neighboring market stations have already been considered. When the single central transmitter is replaced by an SFN, such separation distances likely will be more than those required by the SFN alone.

There also may be cases in which an SFN is designed without regard to any planned single central DTV transmitter. In such a circumstance, the impact of the SFN on neighboring market stations can be found on the basis of the required protection ratios, ERP of the transmitters, and configuration and topology of the SFN. For more precise calculations, the cumulative effect of all SFN transmitters should be taken into account. Based on the results of the calculations, SFN design parameters can be adjusted to avoid unacceptable effect on neighboring market stations. It is expected, however, that network transmitters in an SFN service area replicating the coverage of a single central transmitter can be closer to neighboring market stations than could the single central transmitter.

8.1.2.2.2 Adjacent Channels

When single central transmitters operating on adjacent channel assignments are located in different markets, they do not share an overlapping coverage area and, unlike the case of adjacent channel stations operating within the same market, there would be no interference spot around the transmitters as occurs when they are sited at different locations within the same market (see Section 8.1.2.1). The only limiting factor for adjacent channel single-transmitter stations operating in different markets is the field strength level of each transmitter, calculated according to location and time availability for interference, at the edge of the coverage area of the other station. The field strength from a transmitter should be low enough to sustain the minimum required D/U ratio for other stations at the edges of their coverage areas.

If one of the adjacent channel assignments is operated as an SFN, then the cumulative effect of all SFN transmitters at any point on or within the protected contour of a neighboring market station should be lower than the limit dictated by the minimum required D/U ratio between those stations.

There may be cases in which an SFN is replacing a planned single central DTV transmitter. In this circumstance, analogous to the case with respect to co-channel neighboring market stations discussed above, the impact of the SFN on adjacent channel neighboring market stations will likely be lower than that of the replaced single central transmitter (see Section 8.1.2.2.1). Consequently, the separation distances already considered between the single central transmitter

and its neighboring market adjacent channel stations will be more than that required by SFN transmitters with respect to those stations. This would be especially favorable for an NTSC adjacent channel assignment operating in the neighboring market because a lower impact from an SFN results in a less-impaired quality of the NTSC picture due to interference from the DTV assignment.

8.2 FCC Interference Criteria

The ability to locate broadcast stations so that they serve the communities intended while avoiding unnecessary interference to neighboring stations depends upon channel allotment and station assignment processes embodied in the FCC Rules in the U.S., the Industry Canada regulations in Canada, and similar regimes in other countries. Such regulatory frameworks have been in place since the earliest days of broadcasting. Methods were developed for analog transmission that are insufficient to handle the spectrum crowding that results from allotting two channels to each station — one for analog and one for digital operations. Consequently, new methods have been developed to more accurately predict service and interference than was possible with the schemes used for analog regulation.

8.2.1 NTSC Allocation Historical Method

Analog (NTSC) full-service station allocation criteria are based upon minimum distance separations between stations, channel frequency relationships, the geographic locations of transmitter sites, and the frequency band of operation (VHF or UHF). For example, a UHF station located in the southeastern section of the United States has a minimum distance separation requirement of 205 miles to the nearest co-channel stations, whereas a VHF station located in the same area would have minimum distance separation requirement of 220 miles to the nearest co-channel stations. In the U.S., there are three zones in which different minimum mileage separations apply to the various classes of stations. The zones, in a broad sense, are the Northeastern U.S. (Zone I), the Gulf Coast region (Zone III), and the remainder of the country (Zone II). These relationships are specified in a table in the FCC Rules, and similar separation requirements are specified in the regulations of other administrations.⁶

Besides minimum mileage separation requirements for co-channel relationships (for example, a Channel 9 to a Channel 9), NTSC allocations are concerned with first-adjacent channel relationships (for example, a Channel 9 to a Channel 8 and/or to a Channel 10). UHF stations additionally have “taboo channel” relationships that have associated minimum distance separation requirements. These taboo channel relationships simplistically can be described as those causing an adverse response in a typical TV receiver to certain combinations of received UHF channels, typically where at least one of the signals produces a high received signal level at the receiver. Various combinations can create at the receiver image responses, cross-modulation and inter-modulation products, as well as interference from one receiver to another by virtue of local oscillator radiation.

The minimum distance separation criteria embodied in the FCC Rules are not premised upon interference-free conditions occurring between stations. For example, if two nearby co-channel television stations are operating with maximum permissible facilities that just satisfy the

6. See the tables in FCC Rules, Part 73, Section 73.610(b)(1) for spacing requirements for analog stations and Section 73.623(d)(2) for spacing requirements for new digital stations.

minimum distance criteria, significant areas of interference between the stations are predicted using the propagation model embodied in the FCC curves.

8.2.2 Field Strength Prediction Models

Two primary propagation prediction methods are used to define the coverage and interference areas of television stations: propagation curves such as those in the FCC Rules⁷ and terrain-sensitive propagation models such as the Longley-Rice model. Each is discussed below.

The FCC propagation curves are a family of curves that calculate predicted field strengths at specific distances based upon the input criteria of frequency band of operation (i.e., low-band VHF, high-band VHF, or UHF), the transmitting antenna height above average terrain in a given direction, and the effective radiated power in that direction. These curves are largely based upon empirical data gathered and analyzed in the 1950s and 1960s. The terrain occurring between 3.2 and 16 km (2 and 10 miles) of the transmitter site is used to predict the signal levels throughout a station's entire service area and the interference zone surrounding it. So as long as the terrain over the entire distance of concern is similar to that occurring between 3.2 and 16 km of the transmitter site, the field strengths predicted are likely to be valid. If the terrain in the area of interest departs widely from the terrain occurring between 3.2 and 16 km of the transmitter site, however, either within 3.2 km of the transmitter site or beyond 16 km from it, then non-valid field strengths are likely to be predicted. Hence, in areas such as the mid-west and the south, where the terrain tends to be relatively smooth, the methodology embodied in the FCC propagation curves provides reasonably good results. On the other hand, in areas where the terrain tends to depart more widely, such as in the western parts of the country, the FCC propagation curves may provide questionable results. Similar results can be expected from the application of comparable propagation curves to the terrain of other countries.

The Longley-Rice propagation model [4] is another field strength prediction tool that is considered to be more accurate than the FCC propagation model in many cases. This is because terrain along the entire path is considered in the calculation, not just the terrain occurring between 3.2 and 16 km from the transmitter site. Commonly, the Longley-Rice model is used to calculate the field strength at each point in a universe of periodic points in a defined grid. At each grid point, the terrain is determined along the path from the transmitter site to that grid point, and using the effective radiated power, frequency, and antenna height as principal input parameters, the field strength at the grid point is calculated. A grid of predicted field strengths is calculated that can then be analyzed to determine the predicted coverage and/or interference areas of a station.

The initial Longley-Rice propagation model, then called NBS Tech Note 101, was first published in 1969. The current Longley-Rice model implementation is in the form of a software program that the Commission uses and that is now in Version 1.2.2 (last updated in 1985). Basically, the model computes excess propagation loss, relative to free-space loss, based upon a combination of optical geometry (ray) theory (within the radio horizon) and Fresnel-Kirchhoff knife-edge diffraction methods for predicting loss over round obstacles.

8.2.3 Single-Transmitter DTV Allocation Planning Factors

In the U.S., the FCC has established guidelines for analyzing the coverage and interference for conventional (single-transmitter) NTSC and DTV facilities in a document entitled "Longley-Rice

7. See FCC Rules, Part 73, Section 73.699, Figures 9, 9A, 10, 10A, 10B, and 10C.

Methodology for Evaluating TV Coverage and Interference. OET Bulletin No. 69.” To determine the interference caused to a station to be protected, this OET-69 interference calculation procedure first bounds the protected service area of the subject station to either the NTSC Grade B contour or the DTV noise-limited contour, as appropriate. These contours are calculated using the FCC propagation curves. Within the calculated contours, a square grid is laid out using regularly spaced (typically 1 or 2 km) north-south and east-west lines to establish a matrix of “study cells” in which to conduct analyses of service and interference. The Longley-Rice propagation model is then employed, without consideration of interference from other sources, to determine in which study cells within the bounded area the signal levels from the protected station are predicted to be great enough to provide service. The population residing within each of the study cells predicted to receive service is also calculated. For those same study cells, further calculations are performed to determine if any interference (for example, from nearby co- and/or adjacent channel stations) is predicted to occur.

Within each study cell located within the bounded, FCC-predicted contour, a hypothetical receiver is located with its directional receiving antenna elevated 9.1 meters (30 feet) above ground level and oriented toward the protected (desired) station. The use of an assumed directional receiving antenna permits the use of “antenna discrimination” to reduce the received signal levels arriving from other stations not in the same orientation as the protected station. The DTV planning factors in OET Bulletin No. 69 assume an antenna with 4, 6, and 10 dBd gain, respectively, for low-VHF, high-VHF and UHF receiving antennas. The front-to-back ratios of the receiving antennas are assumed to be 6 dB for NTSC in all cases and 10, 12, and 14 dB, respectively, for DTV low-VHF, high-VHF, and UHF.⁸

To calculate whether a study cell is affected by interference caused by other stations, the Longley-Rice propagation model is again used to predict the received signal level from possible interferers to that cell, considering receiving antenna discrimination. The FCC, in OET-69, also established desired-to-undesired (D/U) signal level ratios, which are used to determine whether a possible interfering station is predicted to cause interference within the cell of the protected station. The D/U ratios used are dependent upon the type of station being protected (DTV or NTSC), the type of interferer (DTV or NTSC), and the channel frequency relationship. If interference is predicted to a study cell, that cell is classified as “masked” and its population counted as receiving interference; other stations then may cause interference to that cell without counting toward increased interference to the protected station.

The FCC has a policy permitting a so-called *de minimis* level of interference to NTSC and DTV stations from any changes made to other DTV and NTSC stations. The *de minimis* threshold is dependent upon how much interference the protected station is already receiving and is calculated using of the aforementioned OET-69 procedure. The permitted *de minimis* level of interference from a DTV facility modification is up to an additional 2 percent of the protected station’s (either DTV or NTSC) population receiving interference-free service (i.e., the total population in cells predicted to receive adequate signal levels and not already predicted to receive interference). An exception is made when a protected station is already receiving 10 percent interference to its service population, in which case no additional interference may be caused. If the protected station is already receiving interference to between 8 and 10 percent of its served population, then the permitted *de minimis* interference is that which will not cause the interference

8. See CEA-CEB6C “TV Receiving Antenna Manufacturers Guide To Categorizing Antennas For Use With The CEA TV Antenna Selector Map Program” for antenna gains that can be expected in the marketplace.

total to exceed the 10 percent limit. For a complete discussion of this procedure, refer to the FCC's Public Notice "Additional Application Processing Guidelines for Digital Television (DTV)".⁹

8.3 Receiver Constraints on Network Design

In an SFN environment, the multiple transmitters effectively create multipath distortions of the signals received by consumer DTV receivers. Since each DTV receiver has limited equalization capabilities (equalization window, relative multipath amplitude, and speed of multipath variation), and those capabilities vary with different generations of receivers, SFN designers must take receiver performance constraints into consideration in the design of networks.

Generally, newer generation receivers have better performance with respect to multipath range, relative multipath amplitude, and speed of multipath variation. An SFN may have a more profound impact on legacy receivers. Tests have shown that receivers manufactured before year 2000 have an equalization window of about $-2\ \mu\text{s}$ to $+20\ \mu\text{s}$ for a single $-6\ \text{dB}$ echo. Receivers manufactured after 2000 but before 2004 have an equalization window of about $-2\ \mu\text{s}$ to $+40\ \mu\text{s}$ — a significant improvement in post-echo performance but not much improvement for pre-echo. This does not mean, however, that a legacy receiver cannot operate in an SFN environment. Usually, pre-echo amplitudes are relatively low. Tests have shown that most legacy receivers can tolerate low-level pre-echoes, on the order of $-10\ \text{dB}$ or lower, with up to a few microseconds range. Meanwhile, use of directional receiving antennas can greatly reduce the effective level of pre-echoes.

Laboratory tests of newer DTV receiver chips and prototypes have shown that contemporary pre-echo performance has significantly improved — to more than $-10\ \mu\text{s}$ for $-6\ \text{dB}$ echoes. ATSC A/74 [2] provides a recommended equalization mask for future receivers as shown in Figure 8.4, in which receivers should be able to process signals having echoes with amplitudes at or below the line at any given time offset. In words, receivers that perform at the recommended level should be able to receive signals with $-2\ \text{dB}$ echoes at up to $-5\ \mu\text{s}$, $-5\ \text{dB}$ echoes at up to $-10\ \mu\text{s}$, $-7.5\ \text{dB}$ echoes at up to $-20\ \mu\text{s}$, $-1\ \text{dB}$ echoes at up to $+5\ \mu\text{s}$, $-2\ \text{dB}$ echoes at up to $+10\ \mu\text{s}$, $-3\ \text{dB}$ echoes at up to $+20\ \mu\text{s}$, and $-6\ \text{dB}$ echoes at up to $+40\ \mu\text{s}$. These multipath performance levels can improve significantly reception in an SFN environment, even when simple receiving antennas are used.

Field tests in an SFN environment also have demonstrated that receiving antenna gain and directivity might not be very important. Receiving antenna steerability is a more important feature for multi-channel reception.

8.4 Managing Network Internal Interference

One of several critical tasks in designing a distributed transmission network is the management of interference between transmitters within the network. Such interference, generally termed "network internal interference" is different from the interference that typically exists between transmitters that are not part of the same network. The difference lies in the facts that the transmitters are all transmitting the same signals and that receivers can be depended upon to treat multiple instances of the same signals at their inputs as echoes of one another. Those echoes result in inter-symbol interference (ISI), but receiver adaptive equalizers are designed to deal with ISI-laden signals and still recover the data they carry.

9. U.S. Federal Communications Commission, "Additional Application Processing Guidelines for Digital Television (DTV)," August 10, 1998, PNMM-8116. [5]

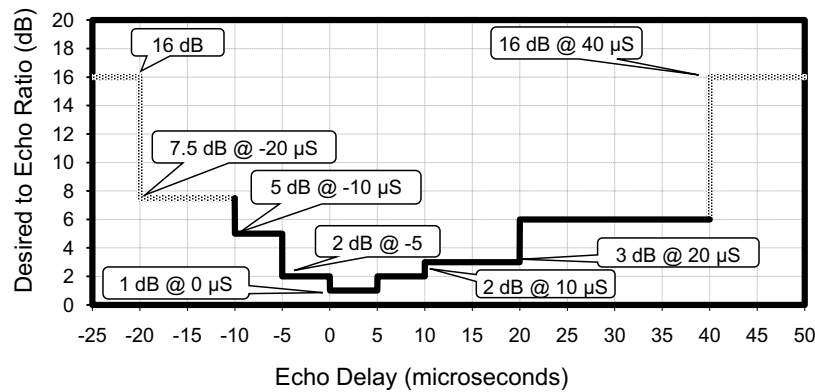


Figure 8.4 Recommended echo delay vs. amplitude capability of future receivers.

There are several tools and techniques available for reducing the impact of network internal interference and thereby making the work of the adaptive equalizer easier. These tools include setting of the spacing between transmitters, control and adjustment of the emission times of the transmitters in the network, and selection of several transmitter parameters with different objectives than would be applied when designing a single-transmitter system.

8.4.1 Transmitter Spacing

One of the most powerful tools available to the designer of a distributed transmission network is the setting of the spacing between transmitters. The spacing between transmitters ultimately determines the differential in arrival times of signals from multiple transmitters at various places within the network. For example, if it is desired to build a network in which the signals from the transmitters are allowed to overlap one another so as to provide the greatest “transmitter diversity” and it is recognized that there is a finite maximum time differential between received signals that receiver adaptive equalizers can handle, then there is a corresponding maximum spacing between transmitters that will enable such operation.

It should be kept in mind that radio waves travel at the speed of light, which equates to approximately $5\frac{1}{3}$ microseconds per mile. If that number is rounded to 5 microseconds per mile for ease of estimation, it becomes relatively easy to think about the impact of transmitter spacing on the difficulty of receiving (synchronized) signals from several transmitters. The difference in the arrival times (in microseconds) of the signals from any combination of transmitters can readily be seen to be five times the difference in miles of the path lengths from the transmitters to the receiver. To cite an example, if a receiver is 5 miles from one transmitter and 10 miles from another, the difference in arrival times of the signals from those two transmitters at that receiver will be about 25 microseconds, assuming that the signals were emitted at the same time. The latter assumption will not be valid in all cases because it often is desirable to adjust the emission times of transmitters in a network, but the concept is still valid. The relative arrival times at the location of the receiver described will be different by 25 microseconds from whatever the relative emission times of the signals were at the two transmitters.

Typically, transmitter spacing, antenna heights, and transmitter power levels are all considered together as part of the design process. When transmitters are spaced relatively close to one another, the objective usually is to provide shorter delay spreads between signals from multiple transmitters. In such a situation it would be counterproductive to operate such transmitters with

high antenna elevations and with high power levels because too many signals would overlap at too long a distance. Thus, for relatively closely spaced transmitters, relatively lower antenna elevations and lower power levels are used. These are generally considered to be “small cell” systems. Conversely, when transmitters are relatively far apart from one another, the objective usually is to reduce the areas of signal overlap from the multiple transmitters in a network and then to manage the arrival times of the signals and other parameters so as to minimize the stress on receiver adaptive equalizers. These are generally considered to be “large cell” systems. Systems can be built that use mixtures of large cell and small cell techniques, but they will be treated separately in this discussion to make clearer the various considerations in their use.

Where broadcasters have existing facilities with antennas on tall towers and high power transmitters and choose to add or convert to SFN operations, they most likely will start with large cell networks and perhaps evolve toward small cell systems. If this process is carried sufficiently far, the high antenna and high power required to provide service to a large area can be replaced by the use of several transmitters with low antennas and power levels to provide more uniform signal levels throughout the service area, potentially at lower cost.

8.4.2 Delay Spread (Timing of Synchronized Transmitters)

In a small cell system, the objective generally is to position the transmitters sufficiently close to one another that the delay spread from all the signals reaching any location in the network's service area will not exceed the maximum delay spread target for the network. To achieve this type of time relationship in a small cell network, the emission times of all the transmitters generally are the same. The transmitters are then positioned in a more or less regular pattern so that the entire area to be served receives signals from at least one transmitter and preferably from several. Since the arrival times of all the signals at all locations in such a network fall within the time windows of receiver adaptive equalizers, advantage can be taken of the opportunities that distributed transmission offers for transmitter diversity to help in reception within homes, in high rise office and residential areas, in big city high rise canyons, and perhaps in pedestrian and mobile applications of the future.

When a large cell system is deployed, there is no inherent assurance that the signal arrival times from multiple transmitters will fall within the time windows of receiver adaptive equalizers. Instead, more complex design and signal management methods must be used to maximize the occurrence of locations where signals can be easily received and to minimize those locations where additional effort is required to receive the signals despite network internal interference. In such networks, advantage is taken of the fact that receiver adaptive equalizers operate on echo signal levels down to some 20 dB below the composite received signal level, treating components at lower levels, even though coherent with the desired signals, in the same way as other noise in the channel. Thus, a C/I ratio is established (generally taken as 20 dB for 8T-VSB systems) above which one signal is considered dominant and the lower level (echo) signals from other transmitters are treated as noise. Where one signal is dominant in this way, the characteristics of the receiver adaptive equalizer are immaterial from a network design point of view. The design effort then becomes focused on managing the network internal interference in those areas where the C/I ratio is predicted to fall below the 20 dB threshold. Techniques that can be used in such locations are emission time offsets of individual transmitters, transmitter power and antenna height, and antenna pattern choices that minimize the areas where C/I ratios do fall below the 20 dB threshold, and putting those areas where there is no significant population.

8.4.3 Emission Timing Adjustments

Among the techniques available for minimizing the effect of signals from multiple transmitters, primarily in large cell networks, is the adjustment of the emission timing of the several transmitters in the network. Emission timing is available as a tool when distributed transmitters and/or distributed translators are used. Emission timing is important in those areas where the C/I ratio between network transmitters falls below the threshold value of 20 dB at which the adaptive equalizers in receivers begin to come into play. Emission timing adjustments result in moving the line that defines the equal time of arrival of signals from two transmitters and the locations of associated differential arrival time values. The equal-arrival-time line is a line perpendicular to the line between two transmitters that, assuming that the two transmitters emit their signals at the same time, intersects the line between the transmitters at its mid-point. On either side of the equal-arrival-time line, signals from the two transmitters arrive slightly offset in time from one another, and loci of constant arrival time differentials describe parabolic curves on either side of the equal-arrival-time line, with one transmitter's signal arriving earlier than that from the other transmitter on one side and later on the other side of the equal-arrival-time line.

There are two strategies that can be followed in setting the emission timing: One is to set the emission timing so that all transmitters in a network emit their signals at the same time; this approach generally would be followed in small cell networks and assumes that the cell sizes are roughly uniform. The second strategy, used mostly in large cell cases, is to adjust emission timing to place the equal-arrival-time line roughly across the locus of equal signal levels from the corresponding pair of transmitters. This has the effect of minimizing the creation of apparent leading echoes by the pair of transmitters under consideration. In such a case, it likely also will be an objective to place the locus of equal signal levels in places where there are lesser populations to be served, and, by association, that becomes the objective for the timing adjustment as well. When more than two transmitters are involved in a network, the relative emission timing of all members of the network must be examined together to find the optimum setting.

8.4.4 Transmitter System Design Parameters

System design parameters for individual transmitters in a network are the same classic parameters that designers of transmission systems have always manipulated to obtain specific coverage and interference protection results for the stations they were designing. The parameters involved include the effective radiated power and antenna height, the directional pattern of the antenna in the azimuthal, or horizontal, plane, and the directional pattern of the antenna in the elevation, or vertical, plane. The difference when setting these parameters for transmitters in a network is that a new set of objectives come into play when it becomes the goal to optimize the performance of the overall network as opposed to the performance of a single transmitter.

8.4.4.1 ERP and Antenna Height

In choosing the effective radiated power and the antenna height, the objective for a network transmitter will be to cover just the area desired and as little more than that as possible. This compares to the general case with a single transmitter in which it is the objective to cover as large an area as possible. The antenna pattern also comes into play, as discussed below.

In the cases of transmitters in small cell networks, overlapping of the signals of adjacent transmitters is expected and desired, with an objective that overlaps occur only between the signals from immediately proximate transmitters and not between the signals of transmitters two cells apart from one another. Since generally it is assumed in small cell designs that the emission

times of the transmitters are all the same and it is likely that the cell sizes may be established so as to take advantage of the full range of receiver adaptive equalizer capability, overlapping of signals from widely separated cells would result in delay spreads outside the range of the adaptive equalizers. Hence, it is necessary to control the regions of overlap to just those between contiguous transmitters. ERP and antenna height for transmitters in small cell networks must be set with this goal in mind.

In the cases of transmitters in large cell networks, the spacing of the transmitters is such that overlapping signals, even from immediately adjoining transmitters, could fall outside the delay spread capability of receiver adaptive equalizers. In these situations, it becomes desirable to minimize the areas of overlap and to transition from one dominant signal to another dominant signal in as short a distance as possible. This minimizes the sizes of the regions in which the C/I ratios between contiguous transmitters fall below 20 dB and thereby require the operation of receiver adaptive equalizers to deal with multiple strong signals. At the same time, minimizing the sizes of the regions falling below 20 dB also has the effect of reducing the maximum differential arrival times to which receiver adaptive equalizers will be subjected, thereby reducing the stress placed upon them. ERP and antenna height for transmitters in large cell networks must be set with this goal in mind.

8.4.4.2 Antenna Design Parameters

Degrees of freedom in antenna design that can be brought to bear in meeting the objectives for both small and large cell applications are the azimuth pattern and the elevation pattern, the latter of which includes the amount of electrical beam tilt, the shape of the pattern outside the main beam, and the like.

8.4.4.2.1 Directional Pattern

The azimuth pattern of the antenna for a cell will generally follow the layout of the cells in the network. Thus, for a square grid of cells, for example, especially in a small cell network, it will be optimal to have nearly square antenna azimuth patterns. This approach allows putting more signal power in the directions in which the cell transmitter must provide coverage at a greater distance, and it reduces signal levels in the directions toward adjacent network transmitters, thereby controlling the amount of overlap. The more likely offset grid, which results in hexagonal cells, has similar considerations, but they can be approximated with omnidirectional patterns.

In large cell networks, often the area to be served by a transmitter may be determined by terrain features, perhaps with a given transmitter serving a particular valley, for instance. In these cases, the shape of the area to be covered will be determined by the terrain, and it may be preferable to choose an antenna azimuth pattern matching the shape of the terrain feature (e.g., the valley), rather than the relative positioning of the transmitters with respect to one another.

Another factor that can dramatically affect the choice of azimuth pattern for a particular cell is its location within the overall network. Transmitters in the center of the network may use patterns that are closer to omnidirectional, while transmitters on the periphery of the network may use patterns that aim their signals into the network service area and minimize signals outside that service area. This allows the transmitters to be placed closer to the outer boundary of the service area without causing increased interference to neighboring stations.

8.4.4.2.2 Optimal Beam Tilt and Elevation Patterns

The optimal beam tilt for antennas in SFNs is likely to be considerably greater than for those used with single transmitters. In the single transmitter case, generally the objective is to project as much power as possible toward the outer reaches of the station's service area. Typically this entails placing the antenna as high as possible and calculating the depression angle to the radio horizon or some other target area. Then the beam tilt is set so that the 90 percent field value on the top of the beam just matches the depression angle for the desired target area or distance.

For SFN transmitters, the objective is to provide service in the region around the transmitter, and other transmitters generally provide the signals for distant locations. The SFN antennas will normally be at lower elevations than in the single transmitter situation, but even taking that factor into account, greater depression angles are usually optimal. This results from the facts that, when small cells are involved, it is desired to keep the signal from reaching beyond the first adjacent cell in the network, and when large cells are involved, it is desired to minimize the size of the region where the C/I ratio falls below 20 dB.

The large cell case also calls for a quick drop-off in the signal level above the peak of the beam and control of minor lobes above the peak of the beam that would otherwise radiate into the territory intended to be covered by an adjacent transmitter. This is contrary to current practice in which most attention is paid to the filling of nulls below the main beam and almost no attention is paid to the pattern above the main beam. Small cell systems also may benefit from attention being paid to the pattern above the main beam.

Yet another factor that can be important in the design of antennas for SFN transmitters relates to the D/U ratio with respect to other stations on adjacent channels in the same market (or in overlap areas between markets). Normally, placing a transmitter on an adjacent channel at a moderate distance from its adjacent channel neighbor results in an area surrounding the transmitter in which the required D/U ratio between the two signals cannot be provided. This is why it is generally best to operate adjacent channel stations in a market co-located with one another. In such cases, because of the close frequency relationship of the two channels and the corresponding likelihood of approximately equal propagation conditions on them, the D/U ratio established at the transmitter more or less will be maintained throughout the service area (assuming that the two stations use comparable antenna patterns).

For the same reasons, operation of an SFN with a single transmitter station on an adjacent channel in the same market may be precluded. A solution that avoids preclusion of such operations is the use of SFN techniques by both adjacent channel stations and the co-location of their transmitters. This works if the two (or more) stations can cooperate, and it can even lead to savings for both stations if they choose to build their transmitter networks together, taking advantage of the many economies of scale that can result.

If there is not such cooperation between the stations, it may still be possible to build an SFN on an adjacent channel if the antennas of the cells can be mounted sufficiently high, the transmitter power levels are sufficiently low, and antenna patterns can be devised that dramatically reduce the power radiated below the antenna so that the necessary D/U ratio can be obtained even in the region around the SFN transmitter. It should be noted that the signal level that can be radiated below such an antenna decreases as the SFN transmitter location is moved farther from the adjacent channel, single transmitter neighbor. Antennas of the sort described are in development, but success in their design is not assured as of this writing. In the future, however, they may become a tool in the kit of the network designer.

8.5 Evolving Network Designs

As time passes, DTx system operators may wish to modify or enhance their networks. The network design and planning processes should consider such future expansion possibilities to facilitate their implementation.

8.5.1 Long Term

Long term planning may be related to the expanding receiver base. At the outset, a network may be designed to cover the bulk of a service area but with some unserved or underserved areas. As the number of receivers increases with time, it later may become economically feasible in some instances to add service to these areas by adding new transmitters or by increasing the power or changing the antenna patterns of existing transmitters. In other instances, it may become feasible to move from large cells to smaller cells in order better to accommodate new services and new receiver capabilities that may appear in the marketplace. Initial planning should consider the effects of and options for such expansion.

8.5.2 Variable Network Configurations

A possibility that exists with DTx systems that is not available with conventional, single transmitter approaches is the customization of network settings to favor different areas at different times. It may be desirable for DTx system operators to change their system settings as a function of time of day or week to optimize coverage. For example, during business hours, system parameters might be set to optimize urban area reception. During evening “prime time” hours, parameters might be adjusted instead to optimize reception in suburban areas. Such variable configurations might also be adjusted as a function of weekday or weekend days or even during special events (such as a major sporting or cultural event).

9 IMPLEMENTING DISTRIBUTED TRANSMISSION SYSTEMS

Distributed Transmission (DTx) systems include those involving use of Distributed Transmitters (DTxTs) and of Distributed Translators (DTxRs). A critical element in the implementation of a distributed transmission system is synchronization of the emissions of each transmitter in the network. This function is accomplished using the techniques and protocols specified in ATSC A/110 [3]. Specific implementation issues relating to A/110 are discussed in the following sections.

9.1 Use of the ATSC Synchronization Standard for Distributed Transmission A/110

A/110 defines a standard for synchronization of multiple transmitters and/or translators emitting trellis-coded 8-VSB signals in accordance with ATSC A/53 Part 2:2007. The emitted signals from such transmitters and/or translators operated according to the A/110 standard comply fully with the requirements of ATSC A/53. In the following discussion, operation of Distributed Transmitters will be considered first with the similarities and differences of Distributed Translators treated in a separate subsection. In most instances, what is described for the case of DTxTs also applies to DTxRs.

A/110 specifies mechanisms necessary to transmit synchronization signals to the several transmitters using a dedicated Packet Identifier (PID) value, including the formatting of packets associated with that PID and without altering the signal format emitted from the transmitters. It also provides for adjustment of transmitter timing and other characteristics through additional information carried in the specified packet structure. The techniques defined for synchronization of multiple transmitters may also be applied to single transmitters when it is necessary or

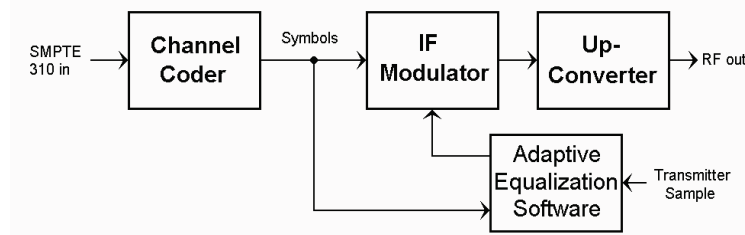


Figure 9.1 Basic ATSC modulator.

desirable to synchronize processes conducted at the input end of a transport link with those carried out at a transmitter location, as may be useful for enhanced transmission technique.

9.1.1 Distributed Transmitter System Hardware Architecture

A Distributed Transmission (DTx) system involves two major elements: a Distributed Transmission Adapter (DTxA) that pre-processes the signals fed to multiple transmitters via an STL (for DTxTs) or over the air (for DTxRs), and Slave Exciters that respond to synchronization and control information placed in the MPEG data stream by the DTxA. The information sent by the DTxA to the Slave Exciters is placed in precursor packets specifically inserted in the data stream just for that purpose by a station's service multiplexer. The MPEG data stream clock frequency is specified in ATSC A/53 as 19.392658...MHz and is subject to the normal MPEG clock frequency tolerance of ± 2.8 ppm. There is an SMPTE standard interface (SMPTE 310M) for transmitters and equipment that feeds transmitters, such as STLs, which provides both an interface protocol and a rate-of-change requirement for the MPEG clock frequency of 0.028 ppm/second. The SMPTE 310M interface may not be used in some applications, but the clock frequency rate of change requirement is still relevant, especially for Distributed Transmission systems as defined in ATSC A/110.

The DTxA and the slave exciter may be derived from an existing exciter design. A block diagram of an existing exciter design is shown in simplified form in Figure 9.1. Basically, an ordinary ATSC exciter consists of a channel coder, an Intermediate Frequency (IF) modulator, and an up-converter. The channel coder performs the various bit manipulations: randomization, Reed-Solomon forward error correction coding, data interleaving, trellis coding, segment and field sync insertion, etc. Its output is a series of ATSC symbols. The IF modulator takes the symbols, performs 8-VSB modulation, and generates a modulated 8-VSB signal at the IF. It may also apply linear and/or nonlinear equalization to correct for downstream distortions in the transmitter. The up-converter does a frequency translation to the operating channel.

A DTxA is formed by deleting the IF modulator and up-converter functions from the exciter and adding some new functions to the channel coder. These new functions are shown in Figure 9.2 and include:

- Distributed Transmission Packet (DTxP) generation and insertion
- Cadence Sync insertion
- Field Rate Side Channel insertion
- Buffering and frequency smoothing of SMPTE 310 input (MPEG data stream)

A slave modulator is formed by taking a conventional modulator and adding some new functions to its channel coder. Specifically, as shown in Figure 9.3, these include:

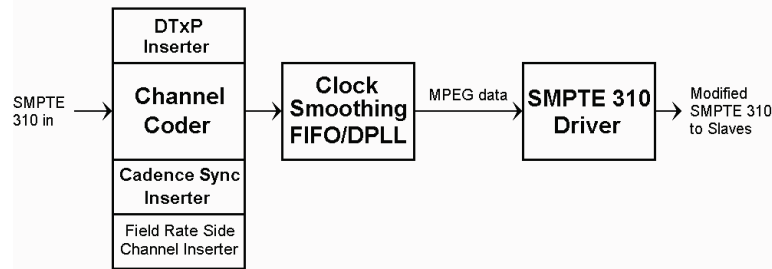


Figure 9.2 Distributed transmission adaptor.

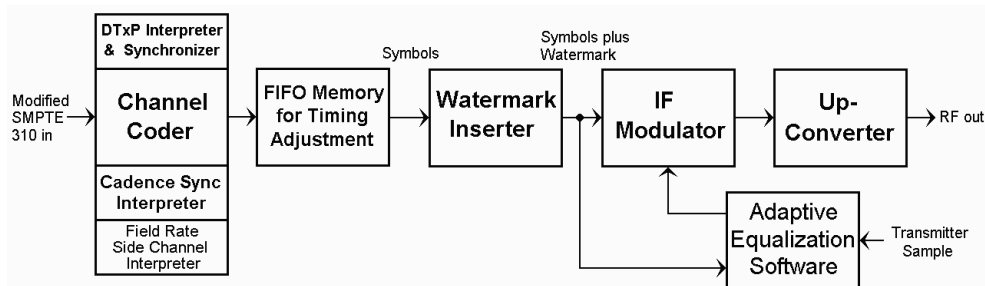


Figure 9.3 Distributed transmission slave.

- DTxP processing and synchronization based on DTxP contents
- Timing adjustment based on DTxP data and cadence sync
- Recovery of field sync reserved bits from field rate side channel
- RF watermark insertion

The channel coding process is best implemented as a “push” design rather than a “pull” design. In a “pull” design, as shown in Figure 9.4, the master process is on the output side of the design, where the MPEG data is multiplexed with field sync, segment sync, Reed-Solomon error correction data, etc. A “pull” design is appropriate for a non-DTx modulator because the insertion point for frame sync is arbitrary.

But in a DTx system, where frame sync timing is not arbitrary, it is crucial to insert sync at the correct time, and to delay the data with the proper timing. Pointers to frame sync in the MPEG bit stream must be maintained through the channel coding process. This requirement is best addressed by using a “push” design, where the logic on the input side of the FIFO memory controls the assembly of the symbol stream at the output, as shown in Figure 9.5.

9.1.2 Frequency Control

For a distributed transmission system to work properly, it is important to control the pilot frequencies of the various slave transmitters so that no two transmitters are more than 1 Hz apart. In other words, frequency control must be accurate to within 0.5 Hz at the transmitter output frequency as measured with respect to the GPS 10 MHz frequency reference. By keeping frequency differences to 1 Hz or less, current consumer receivers will be able to lock on to what appears to be slowly varying multipath (Doppler shift).

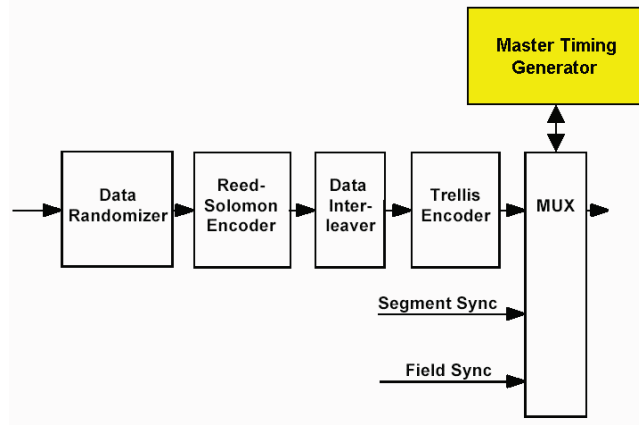


Figure 9.4 A simplified “pull” design.

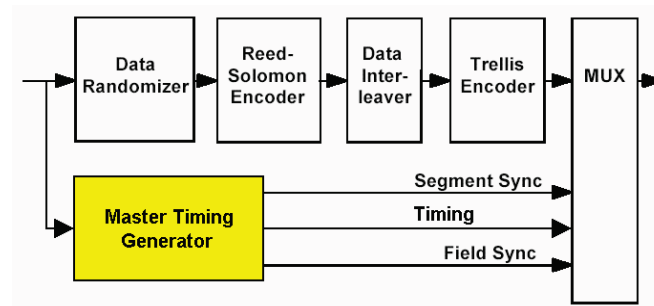


Figure 9.5 A simplified “push” design.

There are two subsystems that determine the pilot frequency of an ATSC signal: the IF modulator, and the up-converter.

It is a straightforward exercise to phase lock all up-conversion oscillators to an external 10 MHz reference taken from a GPS or rubidium source. Such an up-converter, capable of locking to an external frequency reference, is a requirement in a distributed transmission system. The frequency error introduced by the up-converter is essentially zero.

This leaves the IF modulator as the only possible source of frequency error. Modulator frequency error in turn has two components — steady state and transient. Most DTV modulators incorporate some kind of frequency correction so that, if the incoming MPEG data stream clock is off frequency, the frequency error does not introduce a steady state error in the IF pilot frequency. A transient frequency error may exist, however, when the MPEG data stream clock frequency is slewing. The design emphasis for a DTx system is to minimize the transient frequency error.

The extent of the IF modulator output frequency error will depend on the architecture of the IF modulator.

When there is an error in the MPEG data stream clock frequency, some parts of the IF spectrum will change frequency, regardless of the architecture of the IF modulator. Since the spacing between the pilot frequency and the Nyquist frequency is exactly one-half of the ATSC symbol rate, and since the ATSC symbol rate is always exactly 564/313 times the MPEG data

stream clock rate, then something in the IF output spectrum will necessarily move when the MPEG data stream clock frequency moves.

In response to an MPEG data stream clock frequency error, some IF modulators keep the pilot at the same frequency and allow the other parts of the spectrum to move. Some IF modulators move the entire spectrum when the MPEG data stream clock frequency changes. Other ATSC modulators keep the center of the IF signal at the proper frequency and move everything else. Some IF modulators correct for pilot frequency errors — but only after a measurement period has passed.

There are two requirements for frequency control of DTx systems embodied in the A/110 standard. First, the DTxA is required to buffer frequency changes in its MPEG data stream input such that its output frequency changes no faster than 0.028 ppm per second (as required by SMPTE 310M) in response to an input discontinuity of 5.6 ppm (representing a change in its MPEG input stream clock frequency from -2.8 ppm to $+2.8$ ppm — the limits of the MPEG clock frequency tolerance band). This requirement may be fulfilled by using a FIFO buffer memory and a slow, second order, type-two, critically damped, phase locked loop, with a time constant of at least 200 seconds.

The second requirement is that slave modulators must keep their pilots within 0.5 Hz of nominal in response to static MPEG data stream clock frequency errors of up to 2.8 ppm and dynamic frequency errors of up to 0.028 ppm/second. If a modulator architecture is such that it inherently does not change its pilot frequency in response to MPEG data stream clock frequency errors, then nothing need be added to its design. On the other hand, if the modulator includes a frequency correction system to keep its pilot on frequency in the presence of static MPEG data stream clock frequency errors, such a system may require modification to stay on frequency in the presence of dynamic frequency errors. This, in turn, may require modifying the frequency correction system to have a faster and more accurate response to dynamic frequency errors.

9.2 Recommended Variable System Design Parameters

When designing a DTx system, consideration should be given to the present and future populations of deployed receivers.

9.2.1 Legacy Receivers (by Generation)

The oldest legacy receivers have the greatest limitations in the performance of their adaptive equalizers. These receivers can tolerate only about 3–5 microseconds of pre-echoes and 20 microseconds of post-echoes. The maximum amplitudes of such echoes that legacy receivers can tolerate is a function of the time displacement from the main signal in a Ricean channel, with a maximum amplitude of about -4 dB for small post-echo time displacements, dropping off to lower values for longer time displacements. Pre-echoes must be limited to somewhat lower amplitudes.

These oldest legacy receiver adaptive equalizers would establish restrictive constraints for DTx systems were it not for the fact that most such receivers are being used with directional receiving antennas.

Although DTx network design may presuppose that older legacy receivers are using directional antennas, consideration should be given to locating DTxTs so that directional antennas do not need to be reoriented to receive the DTx signal. In many markets, there exists an “antenna

farm” towards which most receiving antennas will be pointed. If possible, DTx network design should assume that receiving antennas will not be reoriented when tuning to the DTxN.

9.2.2 Current Generation Receivers

Current generation receivers have adaptive equalizer windows as wide as -50 to $+50$ μs , with no significant penalty for pre-echo versus post-echo amplitudes. At small time displacements (up to about 3 μs), 0 dB echoes can be successfully equalized.

9.2.3 Future Receiver Generations

Future receiver designs can be expected to further increase the equalizer window and the range over which 0 dB echoes may be equalized. Other anticipated improvements include the use of E-VSB packets for improved equalizer training and the use of intelligent, electronically steerable, directional receiving antennas. All of these developments will make the design and deployment of DTx systems easier and more effective.

9.3 Utilization of Computer Modeling

One of the important aspects of designing multiple transmitter networks is the use of computer modeling to optimize performance of the systems and to minimize interference, both internal and external to the network. The interactions of multiple transmitters with one another and their impacts on and from facilities external to their networks can be quite complex. A variety of analytical methods is required in order properly to design and optimize such systems.

Chief among the required computer models is the propagation analysis tool with interference calculation capabilities. It must have the ability to aggregate the signals from the transmitters in the network in order to evaluate their impacts both upon one another and on the signals of transmitters outside the network. It similarly must have the ability to analyze interference from transmitters outside the network to those within the network. In the latter case, a major requirement is the ability to determine interference to the network without overestimation because of the likelihood that specific areas analyzed could receive signals from multiple transmitters. In such situations, multiple counting of interference must not occur.

Another factor involving propagation analysis tools is the range of propagation models supported. The FCC’s interference determination and processing methods are dependent upon the Longley-Rice method. Other administrations may require the use of other methods such as the Terrain Integrated Rough Earth Model (TIREM) or certain ITU-R methodologies. In any event, the availability of an assortment of propagation analysis methods is to be preferred because of the potential for discovering certain issues with ones of them that may not be exposed with others.

Besides the pure interference analysis problem, computer modeling will be called upon to evaluate expected service, based upon the areas predicted to receive specific ranges of field strengths or of received signal levels, to determine network internal interference, based upon the $C/(I+N)$ values at particular locations and using specific receiving antenna patterns, and to analyze the emission time offsets between transmitters that will minimize the burdens placed on consumer receiver adaptive equalizers. In addition to these required applications, computer programs that can be integrated with the other modeling tools to evaluate link budgets and to support the specifics of designing antenna patterns to aid the design process. Antenna patterns, in particular, are likely to be different in multiple transmitter networks from those used in conventional, high power, high elevation broadcasting systems

9.4 Similarities of Distributed Translator System Hardware Architecture

Distributed Translator (DTxR) systems are similar to Distributed Transmitter systems in some ways and are different in others. The similarities include:

- DTxRs include networks of synchronized transmitters sharing a common channel.
- DTxRs use the DTxP for control.

Differences include:

- DTxRs use off-air signals instead of STLs for delivery of their input signals.
- DTxRs do not use cadence sync or the field rate side channel because they do not exist in off-air signals. DTxRs require a special receiver that recovers field sync reserved bits for E-VSB operation. Cadence sync is replaced through use of the packet number in the DTxP, which is a pointer to cadence sync.
- For anything beyond one tier of distributed translators, a special DTxA is required to process the MPEG data stream through multiple levels of channel coding models.

As far as Baseband Equalization Distributed Translator (EDTxR) systems are concerned, they have similarities and differences with both Distributed Transmitter Systems and Baseband Equalization Digital On-Channel Repeaters (EDOCR). While EDTxR systems are similar to Distributed Transmitter Systems in the way that they too can form networks of synchronized transmitters sharing a common channel, they do not use any DTxP for synchronization and control purposes. Similar to DTxRs, EDTxRs also use off-air signals instead of STLs for their input delivery. Multi-tier operation of EDTxR systems, however, is limited by the amount of error added to the signal while it is passing through the analog channels to reach the last translator.

Concerning similarities between EDTxR and EDOCR, the two systems have basically the same structure in the way of down-converting the signal to baseband, applying equalization and filtering, up-converting and retransmitting again. EDOCR, however, transmits on the same RF channel as its input, but EDTxR transmits on another RF channel. Such a difference in the input and output channel brings the possibility of improving EDTxR implementation and operation, and makes EDTxR different from EDOCR in the following ways:

- The output power limitations of EDOCR do not apply for EDTxR.
- Better filtering and signal processing, involving more delays, can be used in EDTxR. Such implementations are limited for EDOCR because of the critical role of the repeater's internal delay.

Frequency synchronization method for EDTxR may differ from that of EDOCR. In DTxR, synchronization must be achieved between different translators' output frequency, while in EDOCR, output frequency of each repeater must be synchronized with its input with a minimized offset.

10 IMPLEMENTING DIGITAL ON-CHANNEL REPEATER SYSTEMS

As mentioned in earlier sections, there are three types of DOCRs: the RF Amplifier DOCR, the IF Filtering DOCR, and the Baseband Equalization DOCR (EDOCR). This Section discusses the operational requirements and parameters of DOCR systems, including RF emission mask, frequency stability, mathematical model of the output spectrum ripple, and antenna isolation.

10.1 DOCR Requirements

Basic considerations for any DOCR system include the RF emission mask, frequency stability, and output signal spectrum ripple. These issues, and others, are addressed in the following sections.

10.1.1 RF Emission Mask

In each country, a spectrum authority or administration specifies the RF emission mask(s) to govern DTV system emissions. A DOCR, as a transmitter, must meet the emission requirements. Some countries have promulgated more than one emission mask for different transmission power levels or adjacent channel separation distances. For example, there may be a relaxed or non-critical mask for low power transmitters and co-located transmitters and a tight or critical mask for high power transmitters and adjacent channel transmitters having certain separation distances. Since DOCR transmission power levels are generally in the range of a few Watts to several hundred Watts, a relaxed emission mask often may be used. Since an EDOCR regenerates the signal at baseband and re-modulates the signal, sideband spillover is expected to be lower.

10.1.2 Frequency Stability

A DOCR output signal must maintain a certain frequency stability to avoid the generation of dynamic multipath distortion or Doppler effect at the consumer DTV receiver. The frequency stability of a DOCR should be the same as that of a distributed transmitter; i.e., within ± 0.5 Hz.

10.1.3 DOCR Output Signal Spectrum Ripple

One important parameter for a DOCR is the output signal spectrum ripple. The spectrum ripple is caused by DOCR output signal loopback to the receiving antenna (signal fed back from the transmitting to the receiving antenna), where the loopback signal is combined with the received signal and re-amplified by the DOCR. As a result of DOCR internal delay, the loopback signal can be modeled as a delayed echo. Depending on the DOCR implementation, the delay is usually a few tenths of a microsecond to a few microseconds. It is important to limit both DOCR internal delay and loopback signal level since, for consumer DTV receivers, the loopback signal is equivalent to a delayed echo while the internal delay will make the direct signal from the main transmitter appear as a pre-echo at the consumer receiver. To limit the pre-echo impact on consumer receivers, the DOCR internal delay should be limited to a few microseconds. The antenna loopback signal also should be well below -20 dB relative to the received signal level.

10.1.3.1 DCR Loopback Signal and Spectrum Ripple Mathematics Model

For simplicity, assuming there is only one loopback signal present, the channel model, or impulse response, at the input of the DOCR can be reduced to

$$h(t) = \delta(t) + \alpha\delta(t - k)$$

where $\delta(t)$, a Dirac impulse function, represents the normalized DOCR received signal path, and the second path, the loopback signal, has a delay k and an attenuation α , which is related to the DOCR output/input isolation, or antenna isolation. The DOCR received signal $y(t)$ is the convolution of received signal $x(t)$ and the impulse response $h(t)$,

$$y(t) = h(t) \otimes x(t) = x(t) + \alpha x(t - k)$$

The equation above can be expressed in the discrete frequency domain as

$$\begin{aligned} Y(n) &= DFT[y(t)] = \left(1 + \alpha \exp \frac{j2\pi nk}{N}\right) X(n) \\ &= X(n) \left[1 + \alpha \cos \frac{2\pi nk}{N} + j\alpha \sin \frac{2\pi nk}{N}\right] \\ &= A(n)e^{j\theta} \cdot X(n) \end{aligned}$$

where

$$A(n) = \sqrt{1 + \alpha^2 + 2\alpha \cos \frac{2\pi nk}{N}}$$

and

$$\theta(n) = \arctan \frac{\alpha \sin \frac{2\pi nk}{N}}{1 + \alpha \cos \frac{2\pi nk}{N}}$$

$A(n)$ is the envelope of the spectrum $Y(n)$. Its first two terms $1 + \alpha^2$ represent the average signal power, where 1 is the received signal power, and α^2 is the loopback signal power. The spectrum ripple peak-to-peak value is $2a$. Therefore, the relationship between output spectrum ripple and the loopback signal level can be calculated as

$$\text{in-band ripple (dB)} = 20 \log [(1 + a)/(1 - a)] \approx 20 \log (1 + 2a)$$

where a is the loopback signal level, and $2a$ is the peak-to-peak spectrum ripple. The equation above can also be rearranged as

$$20 \log (a) = 20 \log \{0.5 \times \{10^{[\text{in-band ripple (dB)/20}] - 1}\} \} \text{ dB}$$

where $20 \log (a)$ is the loopback signal level power relative to the received signal power. For different output spectrum ripple values, the tolerable loopback signal levels are presented in Table 10.1. It can be seen that for 1 dB peak-to-peak ripple, the loopback signal must be 24.3 dB below the received signal.

Table 10.1 OCR Output Spectrum Ripple and Loopback Signal Levels

DOCR Output Spectrum Ripple (dB), (peak-to-peak)	0.1 dB	0.25 dB	0.5 dB	0.75 dB	1 dB
Loopback Signal Level 20 log (a) (dB), (relative to received signal)	-44.7 dB	-36.7 dB	-30.6 dB	-26.9 dB	-24.3 dB

10.1.4 DOCR Design Parameters

This section addresses key design parameters of a DOCR system including RF or IF processing and the advantages of the EDOCR approach.

10.1.4.1 RF and IF Processing DOCRs

There are two important parameters governing DOCR operation: the received signal power, and the transmitting-to-receiving antenna isolation, since they decide the output signal power level and the spectrum ripple.

RF and IF processing DOCRs are equivalent to an amplifier or can be modeled as a linear device. The relationship between the antenna isolation, spectrum ripple, and allowed amplifier gain is

$$\text{amplifier gain} = \text{antenna isolation} + 20 \log(a)$$

or

$$(\text{output signal power}) / (\text{received signal power}) = \text{antenna isolation} + 20 \log(a)$$

where $20 \log(a)$ is the relative loopback signal power, as found in Section 10.1.3.1.

For example, if the receiver input signal power is -40 dBm, antenna isolation is 104.3 dB (a value that is not very difficult to achieve), allowed spectrum ripple is 1 dB, or $20 \log(a) = -24.3$ dB, (see Table 10.1), then, the DOCR transmitter power output is limited to

$$-40 \text{ dBm} + 104.3 - 24.3 = +40 \text{ dBm, or } 10\text{W}$$

In this case, if the transmitter power output is higher than 10W , the spectrum ripple will be larger than 1 dB. If higher output power is desired, the antenna isolation or received signal power must be increased.

10.1.4.2 EDOCR and its Advantages

In an EDOCR, a baseband equalizer is implemented to correct the loopback signal. It is equivalent to increasing the antenna isolation through use of digital signal processing. By implementing a baseband equalizer, a -4 dB loopback signal (relative to the received signal) can be corrected. Assuming 1 dB spectrum ripple is allowed, in comparison to the RF and IF DOCRs example, there will be $-4 - (-24.3) = 20.3$ dB of additional margin. The additional margin can be applied to reducing the antenna isolation requirement for easy installation, increasing the output power level, decreasing the required input signal level, or a combination of these. For example, if the received signal is -40 dBm, and, for safe operation, a safety margin of 6 dB is used, the EDOCR transmitter power output will be

$$-40\text{dBm} + 104.3 - 4 - 6 = +54.3\text{dBm, or } 269\text{W}$$

This example shows the advantage of the EDOCR approach. It provides much more flexibility with respect to the received signal level, output power level, and antenna isolation.

Since an EDOCR internally cleans up and regenerates the 8-level baseband signal, it can also provide better spectrum shaping to comply with RF emission mask requirements. First adjacent

channel EDOCR operation is viable, which is almost impossible for an RF processing DOCR and very difficult for an IF processing DOCR. An EDOCR can also provide higher output S/N.

10.1.5 EDOCR RF Watermark Insertion

Since an EDOCR regenerates the baseband signal, it can erase the RF Watermark carried by the main DTV transmitter signal and replace it with another RF Watermark. RF and IF processing DOCRs have to carry the same RF Watermark as transmitted by the main transmitter. This means that an EDOCR can be treated as a distributed transmitter since it can transmit relatively higher power (in comparison to RF and IF DOCRs) and can have the options of either reinserting the RF Watermark received from the main transmitter or inserting its own RF Watermark in its output signal.

10.2 Recommended DOCR System Design Parameters and Receiver Issues

Based on the examples and discussions in the previous section, DOCR system design and operational parameters are listed in Tables 10.2 and 10.3 for RF and IF DOCRs and for EDOCRs, respectively.

Table 10.2 RF and IF Processing DOCR System Design and Operational Parameters

RF and IF DOCR System Design and Operation Parameters:	
Antenna isolation	> 105 dB
Recovered signal power	> -50 dBm
Output spectrum ripple	< 1 dB (internal multipath signal level < -25 dB)
Frequency stability	+/- 0.5 Hz
Output power	< 20W
Internal delay	< 1 μ S
Output S/N	> 27 dB
RF Watermark	Can only pass on the received RF Watermark
First adjacent channel operation	RF DOCR: "No"; IF DOCR: "Possible"

Table 10.3 Basedband Equalization DOCR System Design and Operational Parameters

EDOCR System Design and Operation Parameters:	
Antenna isolation	> 90 dB
Recovered signal power	> -65 dBm
Output spectrum ripple	< 1 dB (feedback signal tolerance level < -4 dB)
Frequency stability	+/- 0.5 Hz
Output power	< 10 kW
Internal delay	< 5 μ S
Output S/N	> 30 dB
RF Watermark	Can create its own or reinsert received RF Watermark
First adjacent channel operation	Yes

The advantages and the flexibility of an EDOCR over other types of DOCR are quite obvious. The EDOCR requires less antenna isolation and less received signal strength, which gives more flexibility in the implementation and possibly reduces the implementation cost (lower cost antennas, easier location selection, reduced installation and antenna isolation cost, etc.). It can

provide higher output power, better spectrum shaping, and/or better output S/N. First adjacent channel operation is viable, which is very important for frequency congested areas (i.e., all the major markets). The EDOCR can also insert its own RF Watermark.

In comparison to large scale distributed transmitter networks, DOCRs may cost less to implement. There is no STL link required and no need to transmit a DTxP. A well-designed DOCR system might not generate very long multipath delay spread, which is friendly to consumer DTV receivers.

10.2.1 DOCR Impact on Consumer DTV Receivers

As described in Section 8, DOCR internal delay causes the signal from the main transmitter to act as a pre-echo in the DOCR coverage area. This result may impact ATSC legacy receivers. Generally, receivers manufactured before year 2000 have an equalization window of about $-2\ \mu\text{s}$ to $+20\ \mu\text{s}$ for single $-6\ \text{dB}$ echoes. Receivers manufactured after 2000 but before 2004 have an equalization window of about $-2\ \mu\text{s}$ to $+40\ \mu\text{s}$ — a significant improvement in post-echo performance, but not much improvement for pre-echo. This does not mean, however, that a legacy receiver cannot operate in a DOCR coverage area. Usually, the pre-echo amplitude is relatively low. Tests have shown that most legacy receivers can tolerate low-level pre-echoes, perhaps $-10\ \text{dB}$ or lower, up to a few microseconds range. Meanwhile, the use of directional receiving antennas can greatly reduce the levels of pre-echoes at receiver inputs. Other good news for DOCR implementation is that a well-designed DOCR system (see Section 8) should experience much shorter post-echoes relative to those that may be generated in large scale distributed transmitter systems.

More recent (i.e., ca. 2004) laboratory tests on new receiver chips and prototypes have shown that pre-echo performance has been significantly improved to more than $-10\ \mu\text{s}$ for $-6\ \text{dB}$ echo. The ATSC A/74 [2] recommended equalization mask for new receivers is $-2\ \text{dB}$ echo at $-5\ \mu\text{s}$; $-5\ \text{dB}$ echo at $-10\ \mu\text{s}$; and $-7.5\ \text{dB}$ echo at $-20\ \mu\text{s}$. This level of pre-echo performance can easily accommodate pre-echoes caused by DOCR internal delay.

10.3 Utilization of Computer Modeling Including

The coverage prediction models for DOCR systems should be the same as those used for distributed transmission systems. No special treatment is needed for DOCR systems.

11 DISTRIBUTED TRANSMISSION SYSTEM DESIGN PROCESS

The design process for a Distributed Transmission Network is iterative in nature. There are a large number of parameters and other characteristics that are variables and must be optimized during system design by repeatedly working through the various stages of design and analysis. The point of optimization will be recognized as improvements in the various performance indicators become more gradual with each successive pass through the process. In the discussion that follows, the various stages will be described as though they are carried out once. It should be recognized, however, that each can be repeated as often as seems useful and that they do not need to be taken in the sequence presented in order to achieve performance improvements. Examples of the results of the various stages of a design process such as that described are given in Annex A.

11.1 Transmitter Placement and Parameter Setting

The first step in the process of designing a distributed transmission network is choosing the number and locations of the transmitters and setting their fundamental parameters. A number of factors bear upon these choices. These include:

- The type of cell structure intended; i.e., large cell, small cell, or even micro-cell structure
- Existence of sites with pre-existing facilities that are intended to be incorporated into the network
- Possibility to utilize facilities previously constructed for other services; e.g., taking advantage of cell phone towers that exist in a region and the fiber networks that interconnect them
- Whether the network is being designed as a cooperative effort among several stations to achieve economies of scale
- Whether signals will reach the various transmitters through a separate STL network or through over-the-air relay

In this regard, it should be noted that the economics of the network design can be treated as a major parameter to be optimized, with the capital expenditure and operating costs of each of the location and related parameter choices constantly weighed against the technical performance achieved. Design of an STL system, if used, will be a separate but parallel exercise, the results of which will determine both the practicality and part of the cost of each of the sites considered for inclusion in the network design. Other fundamental parameters that are optimized at this stage include:

- The effective radiated power at each location
- Antenna height
- Transmitter power output
- Basic type of transmitter design; i.e., solid state or tube, number of cabinets, floor space, etc.
- Antenna pattern

11.2 Coverage Prediction, Interference Analysis, and Population Counting

Once a set of locations and site parameters has been determined, that set is evaluated for the coverage it will provide over the intended service area, the interference it will cause to neighboring stations, and the population that will be reached with various signal levels. This can be accomplished using classic computer-based propagation modeling tools. The computer models will have to be set up to study the combination of results from a number of transmitters rather than from a single transmitter as has historically been the case. The studies will include prediction of the areas receiving field strengths determined to be greater than the required threshold values for the types of services to be offered by the network and interference analyses including the signals from all transmitters in the network as they would impact the services of neighboring stations. Finally, the population predicted to be served with signal levels of various categories would be used as an optimization tool to evaluate the predicted technical performance of the network.

When it comes to evaluating interference between a network design and other stations, some changes in procedures are required from hitherto-used calculation methods. The signal levels from all of the transmitters in the network must be aggregated at each study point in the service area of a neighboring station under study. This aggregation must take account of the effects of directional receiving antennas at each study point if such directional characteristics are specified

in the procedures of the regulatory body having jurisdiction over the location where the network is to be built.

Determining where coverage is achieved with different signal levels involves establishing a grid of study points for the entire expected service area of the network and evaluating the field strengths delivered to each of those points by the transmitters in the network. The transmitter with the strongest field strength at each point is taken as providing service there, and its field strength at that point is used to represent the signal level delivered by the network. Once field strengths delivered to all the study points are known, it becomes possible to apply various threshold values to determine whether to predict service at the locations of the study points based upon the types of services to be offered and their required minimum signal levels. Categorization of the types of services can include whether reception with indoor antennas is intended, whether there will be service to pedestrian or mobile receivers, whether there are any transmission services with enhanced robustness, and the like.

When studying interference from other stations to the network, the areas served by all transmitters in the network must be considered together so that interference is evaluated to the network as a whole and not to the signals of individual transmitters. To obtain this result, the previously described grid of study points covering the overall network is used. Then, at each study point, the transmitter having the strongest field strength at that location is selected as the one providing service there, and any directional receiving antenna at the study point is assumed to be pointed toward the selected transmitter. With those conditions having first been established, the analysis becomes a matter of D/U ratio calculations in the normal manner between the signals from two transmitters, with receiving antenna discrimination applied in the normal manner.

Once areas receiving adequate signal levels are determined and areas receiving interference from other stations are identified, the areas receiving service without interference can be ascertained. Standard methods of counting the population represented at each grid point can then be applied to help in understanding the level of service expected and in optimizing the network design.

11.3 Locating Network Internal Interference

In order to manage network internal interference, it is necessary first to know where it will occur, given any particular network layout and set of design parameters. Network internal interference is considered to occur in locations where the strongest signal from a network transmitter is not at least 20 dB stronger than the signals from the other network transmitters in the aggregate. Most propagation modeling tools provide a method for making determinations of the sort described, typically terming them studies of $C/(I+N)$ for the strongest server. In these methods, a study grid is established across the area to be studied, and the field strength from each transmitter to each study point is calculated using one of the standard propagation models such as Longley-Rice, TIREM, and the like. Once all of the field strength values are determined from each transmitter at each study point, it becomes a simple mathematical exercise to rank order them by field strength, select the highest value, and compare it to the composite value for all the others.

It has been found useful to present on a map the results of comparisons of the field strengths of the strongest server at each study point with the aggregated field strengths of the other transmitters. Color coding can be used to indicate those areas with $C/(I+N)$ greater than 20 dB, where effectively one transmitter is dominant and it can be considered that adaptive equalizers are not required to deal with the weaker signals, and those areas having lesser $C/(I+N)$ ratios. Typically the color coding is used to divide the region below 20 dB $C/(I+N)$ into a number of

ranges down to 0 dB $C/(I+N)$. At the 0 dB point, signals of equal amplitude from one or more transmitters can be expected, and bobbing channels may result. Ranges between 20 dB and 0 dB $C/(I+N)$ may present increasing levels of difficulty for adaptive equalizers as the value approaches the 0 dB end of the scale.

Once the $C/(I+N)$ values are mapped, the network design can be examined to determine the populations located in the various ranges. Network optimization can be iterated to maximize the population in areas having dominant transmitters and to minimize the population in areas having low $C/(I+N)$ values. Moving these areas around on the map will involve changes in transmitter locations, antenna heights and patterns, transmitter power levels, and similar characteristics, as described above.

11.4 Planning Transmitter Timing

Mitigation of network internal interference is treated in parallel with optimization of the locations where it occurs. As described elsewhere herein, an optimal configuration places the locus of equal arrival times of signals from multiple transmitters in the vicinity of the areas receiving equal field strengths from the same transmitters. Doing so results in the least impact from leading echoes – the most difficult condition for receiver adaptive equalizers to correct. In small cell systems, such adjustments are minimal since the optimum configuration generally is one with all transmitters emitting their signals at the same times. In large cell systems, however, there may be benefits from displacing the emission times of the several transmitters by significant amounts. As soon as more than two transmitters are involved, the timing relationships that produce the optimal results become complex.

Similar to the case with the locations of network internal interference, it has proved useful to present delay spread values color coded on a map. The delay spread values are divided into ranges from one exceeding the maximum value planned for the system to a range including zero delay spread (i.e., equal arrival times of the signals). Typically the delay spread values are masked by the $C/(I+N)$ values representing network internal interference so that delay spread is presented only at those locations where one transmitter is not dominant and receiver adaptive equalizer operation becomes significant. Generally the threshold used for masking is the 20 dB $C/(I+N)$ value discussed elsewhere.

With the delay spread values presented graphically after masking by $C/(I+N)$, it becomes a relatively straightforward exercise to position the zero delay spread loci where desired. What is more complex is balancing the larger delay spread values on either side of the zero delay spread loci so as to minimize the overall impact on receiver adaptive equalizers. Determining the areas in which leading echoes are predicted can help in understanding likely network operation, but it is not so easily done as simple mapping of time offsets. In the end, results obtained from analyses of the time offsets must be combined with analyses of the network internal interference and population data to predict populations whose receivers will be exposed to various levels of echoes having various ranges of differential arrival times.

12 MEASUREMENT TOOLS FOR DISTRIBUTED TRANSMISSION NETWORKS

Accurate setup and periodic maintenance of a distributed transmission network is important for proper long-term operation. ATSC A/110 [3] introduces the technique of an RF Watermark to support the sorts of measurements required in Distributed Transmission Networks for setup and maintenance. Test and measurement (T&M) equipment designed and/or configured to work with the RF Watermark also is required to support operation of such networks.

12.1 Use of RF Watermark

The RF Watermark is a valuable tool for adjustment and field verification of Distributed Transmission Networks. Proper implementation of the RF Watermark requires understanding of the theory on which it is based and the applications for which it is intended. With this background, other applications for which it also is suited become evident.

RF Watermarks serve two fundamental purposes when carried by ATSC 8-VSB signals: identification of transmitters and measurement of many characteristics of the signals when they are received. This discussion starts with the basic theory of a Pseudo Random Binary Sequence (PRBS) embedded in an 8-VSB signal to identify the Emitting Transmitter and to enable test equipment to perform DTxN measurements and analysis in the field. It then moves on to the application of the embedded PRBS and several methods for its recognition and use in making measurements of various sorts.

Using an RF Watermark, identification and measurement of signals from several transmitters can be accomplished without requiring shutting them down individually in order to determine which transmitter or transmitters are contributing to the signals received at any given location. Moreover, the channel impulse response components received from each transmitter can be determined. This determination can allow observation and in-service system adjustments of such characteristics as power levels and delay offsets.

The terms auto-correlation and cross-correlation are important to the discussion and are explained through an example using a simple Code Generator. Then, the attributes of the Kasami Code Sequence Generator selected for use in the RF Watermark are briefly discussed. Implementation techniques that were chosen to reduce the computational complexity of test equipment, as well as certain techniques that increase the performance of the RF Watermark when used at low injection levels into an 8-VSB host signal, are discussed. The syntax of the Distributed Transmission Packet (DTxP) is examined from the perspective of enabling transmitter identification, with the corresponding possibility of automated DTxN measurements in the field. After covering these basics, the next sections explore application of the RF Watermark to Distributed Transmitters, Distributed Translators, and Digital On-Channel Repeaters, measuring and adjusting emission timing in a DTxN, selecting measurement locations for network adjustments, and field verification of network design parameters.

Also explored is use of the RF Watermark to permit each transmitter simultaneously to broadcast site-specific data, such as status and telemetry, to a DTxN field monitoring or data collection point. Such return channels may be of interest to broadcasters wishing to return data from one or more transmitters without requiring separate telemetry channels for the purpose.

12.1.1 Basic Theory

The same basic concept that allows multiple cell phone users to share a channel, transmitting at the same time to improve channel utilization efficiency, is applied to the RF Watermark, to allow the signals from multiple transmitters sharing the same channel to be identified. In cell phone systems, the technique is termed Direct Sequence Spread Spectrum – Code Division Multiple Access (DSS-CDMA); in the RF Watermark for 8-VSB transmission, the technique is termed Buried Spread Spectrum (BSS). The BSS signal uses techniques similar to those of CDMA applied coherently with the 8-VSB DTV signal.

To consumer DTV receivers, RF Watermark signals appear to be random noise. The pseudo-random noise (PN) sequences used, however, are not really random. They are deterministic

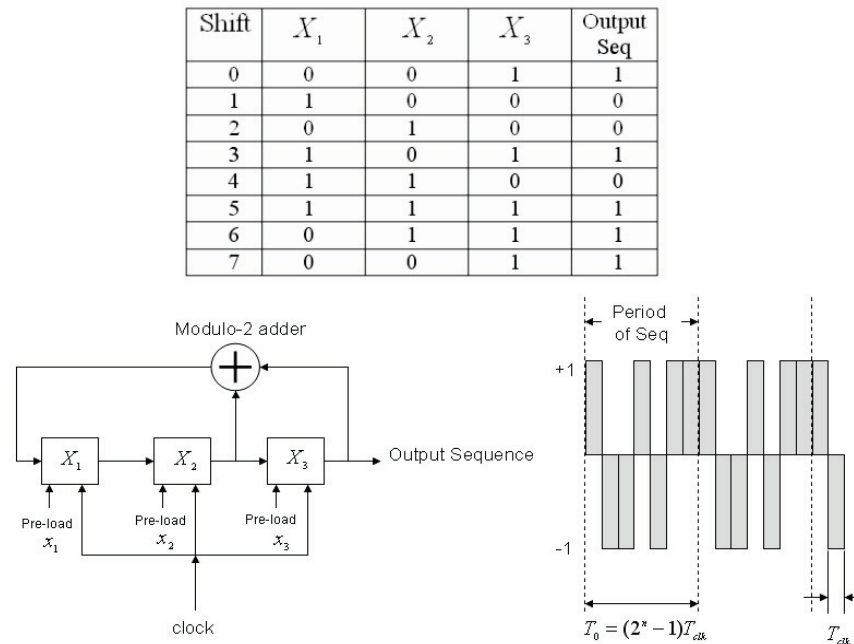


Figure 12.1 Three-stage linear feedback shift register.

periodic sequences that can be detected by special-purpose receivers. The particular form of PN sequences used in the RF Watermark are known as Kasami sequences. They are generated by combining the outputs of several linear feedback shift registers. A linear feedback shift register (LFSR) consists of several consecutive D flip-flops and feedback logic.

Figure 12.1, shows a simple, three-stage LFSR as an example. A binary sequence is shifted through the shift register in response to clock pulses. The contents of the stages are logically combined and fed back to the input. A feedback shift register and its output sequence are termed linear when the feedback logic consists only of modulo-2 adders. The initial states of the stages can be set through a pre-load input, which, together with the feedback logic, determine the successive contents of the stages and, hence, the output sequence produced. This deterministic sequence will repeat itself after a certain number of clock pulses. The sequence length is equal to $2^n - 1$, where n is the number of stages in the shift register. In this example, a sequence length of 7 is expected before the sequence repeats. The table in Figure 12.1 shows a pre-load at shift zero of 001. When clocked, this produces the output sequence 1001011; the sequence will then repeat in a deterministic manner. Using this simple PN generator and by selecting the pre-load values, seven unique code sequences can be generated. The PN code generator selected for the RF Watermark, as discussed later, has 24 stages and can generate more than 16 million unique sequences.

Because of the need to detect particular code sequences while rejecting others that may be present simultaneously, the cross correlation properties of the codes used for transmitter identification are very important. The receiver of a PN sequence must know the code generator that was used by the transmitter as well as was the pre-load starting state. Then, a Correlator circuit can be used in the receiver to lock onto and identify a particular sequence if it is present. Figure 12.2 shows a correlator used to detect the existence of a particular sequence output from the example three-stage linear feedback shift register. The initial pre-load state, 001, is known,

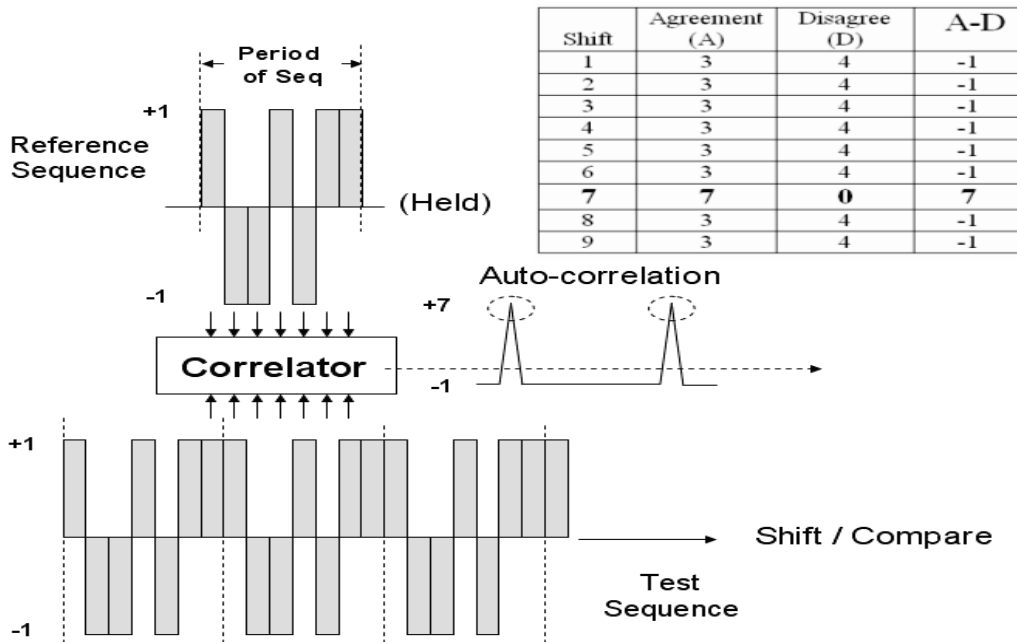


Figure 12.2 The basic correlator circuit.

and the same code generator is used to generate a period of the corresponding sequence, which is called the “Reference Sequence.” The reference sequence is held in memory and is used as one of the inputs to a correlator, and an auto-correlation process is performed on the test sequence.

Auto-correlation refers to the degree of correspondence between a sequence and a phase-shifted replica of itself. Because of the deterministic nature of the identical code generators used, the reference sequence in the receiver will be a replica of the transmitted test sequence, but a phase-shift between the two sequences may be present. The received test sequence is clocked into the correlator, and the correlator samples from each sequence input a number of bits (in this case, 7) equal to the length of the sequence, looking for agreement. The test sequence is shifted into the correlator repeatedly, and all bits of the sample period are checked for agreement with the reference each time the test sequence is clocked. On the clock pulse at which the reference and test sequences come into alignment, the correlator will output a “spike” or auto-correlation peak, which will repeat every code sequence period.

The auto-correlation plot and the table in Figure 12.2 show simply the number of agreements minus the number of disagreements found for the pairs of bits in each shifted sample. When the two code sequences are identical, a correlator produces an output spike once per code sequence period; when they are not the same, no such output spike is produced. This characteristic of auto-correlation is used to great advantage in wireless communication systems today, and is applied in the RF Watermark system.

A Kasami code generator was selected for the RF Watermark because it was found to have a low probability of false synchronization; i.e., good cross-correlation properties. Cross-correlation is the property of a code generator that expresses the extent to which different codes that it generates correspond with one another. Ideally, any two different code sequences from the same generator should have low correspondence or cross-correlation so that multiple codes from a set

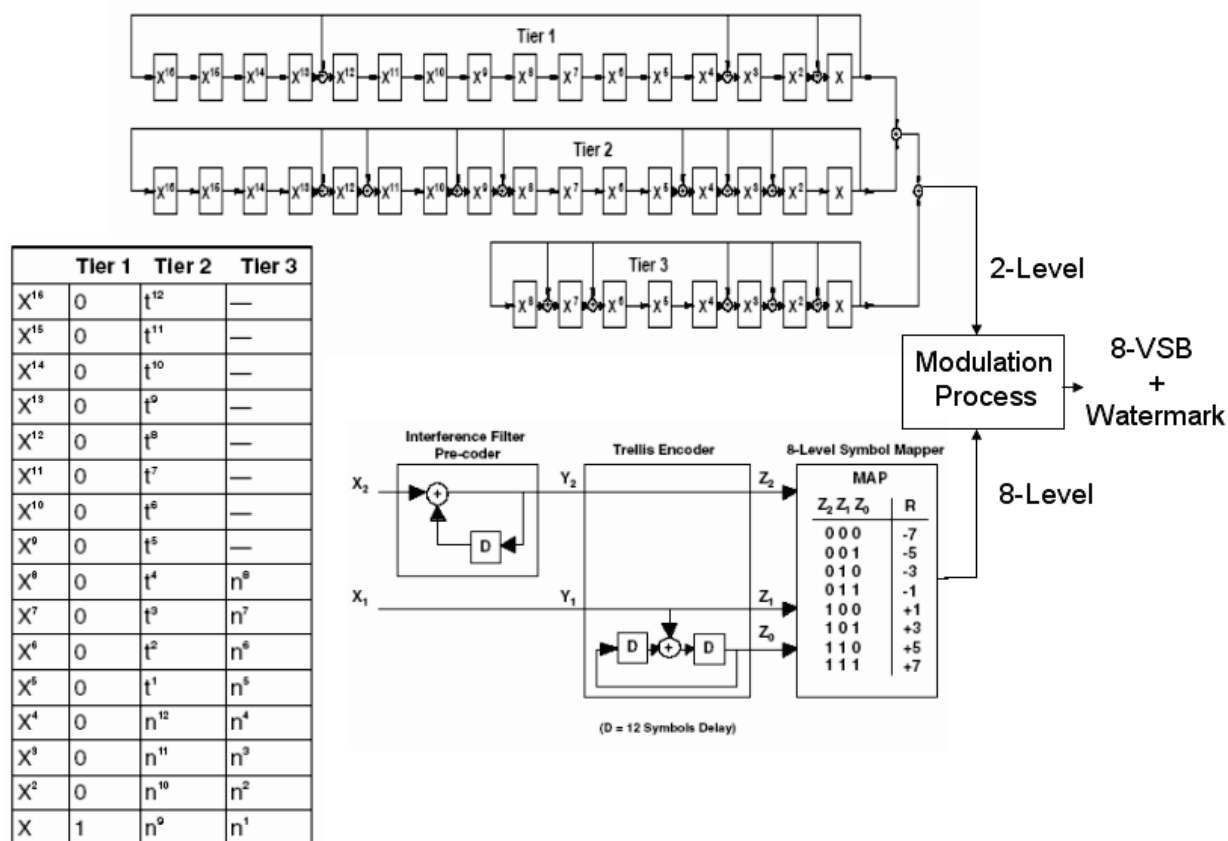


Figure 12.3 Kasami code sequence generator in an 8-VSB exciter.

can be used to enable multiple access to a channel. The truncated Kasami code generator selected for the RF Watermark, exhibits the desired attribute of low cross-correlation.

A Kasami code sequence generator comprises three tiers of linear feedback shift registers, each configured to generate maximum length codes, that are added modulo-2 at their outputs to create a single, combined code. Figure 12.3 shows the configuration of the Kasami code sequence generator that is specified for the RF Watermark in ATSC A/110 [3]. This particular Kasami code sequence generator provides 24 bits of pre-load capability and produces over 16 million unique sequences with good cross-correlation properties.

The table in Figure 12.3 shows the pre-load, or seed, assignments for each of the shift register tiers. Tier #1 is always preloaded with a fixed bit value of 1 in the X position and all the other stages set to zero. The remaining two tiers have a total of 24 pre-load bits that can be assigned by broadcasters. The DTxP carries 12-bit Network Identifier and 12-bit Transmitter Identifier fields. The bits in these fields are assigned as shown in the columns for Tiers 2 and 3 as (n) and (t), respectively. The 12-bit Network Identifier (n) provides a maximum of 4096 codes to identify broadcasters operating a DTxN or single transmitter on a given channel, and the 12-bit Transmitter Identifier (t), allows each broadcaster a maximum of 4096 transmitters for each Network Identifier. A Network Identifier (12 bit) value will be assigned to each network, and a separate 12-bit Transmitter Identifier value will be assigned to each transmitter in a DTxN.

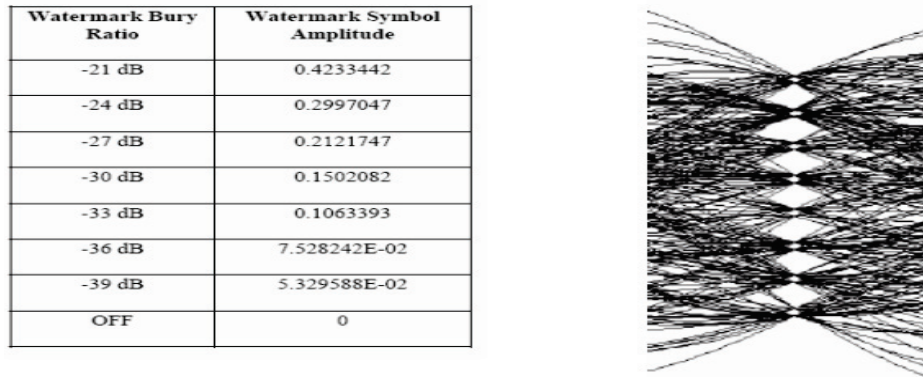


Figure 12.4 RF watermark injection level.

In the RF Watermark system, the Kasami code sequence generator is clocked synchronously with the ATSC symbol clock at 10.76... MHz. The binary Kasami output sequence is used to modulate a parameter determined by the desired bury ratio; i.e., the ratio between the amplitude of the host 8-VSB signal and the amplitude of the RF Watermark. A binary 1 corresponds to +1 weighting and a binary 0 corresponds to -1 weighting of the symbol amplitude values shown in the table in Figure 12.4, corresponding to the various bury ratios defined for the RF Watermark. The resulting 2-VSB RF Watermark symbols are added to the host 8-VSB symbols before they are passed to the digital-to-analog conversion process of the 8-VSB modulator. This produces a low-level binary “RF Watermark” signal riding on and synchronized with the 8-VSB host signal. The bury ratio to be used by each transmitter is controlled through a field in the DTxP; -30 dB is suggested as the nominal level. Since the RF Watermark 2-VSB symbols are synchronous with the 8-VSB host signal symbol clock, a series of very small, additional eye openings appears in an Eye Diagram of the combined signal, corresponding to the 2-VSB RF Watermark. Eight such eye openings are visible in the diagram of Figure 12.4. Note that this simulation used a -27 dB bury ratio — a stronger than normal RF Watermark level — to make the RF Watermark more visible.

Figure 12.5 shows that, in addition to symbol synchronization, the RF Watermark is also time-synchronized to the data frame of the 8-VSB host signal. The RF Watermark is not injected during Data Field Sync data segments; the pre-loading of the Kasami code sequence generator occurs during those intervals. The Kasami code sequence begins with the first active data segment and continues for 312 total data segments until the next Data Field Sync data segment; the process then repeats. Starting and stopping the RF Watermark can reduce the computational complexity needed in an RF Watermark receiver. By locating the 8-VSB Data Field Sync data segment, the receiver also effectively locates the beginning of a 16-bit Kasami code sequence. The length of the Kasami code sequence is equal to $2^n - 1$; with $n=16$, the length is 65,535 bits. Since the RF Watermark modulation is 2-VSB, it carries 1 bit per symbol. The total number of symbols in 312 segments is 259,584. The Kasami code sequence iterates four times per data field. It runs for 3 full periods of 65,535 bits (and symbols) each; then the fourth period is stopped short at 62,979 bits (and symbols), resulting in a slightly truncated code. Simulations have shown that this truncation has minimal effect on the auto-correlation and cross-correlation properties of the Kasami code sequence.

Figure 12.6 shows two views of a simulation of the autocorrelation of the truncated Kasami code sequence after injection with a bury ratio of -30 dB into an 8-VSB host signal. The same 12-

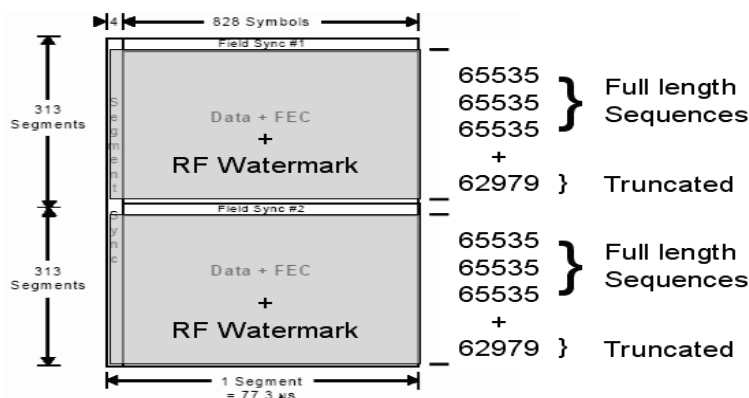


Figure 12.5 Timing of the RF watermark symbols relative to the 8-VSB frame.

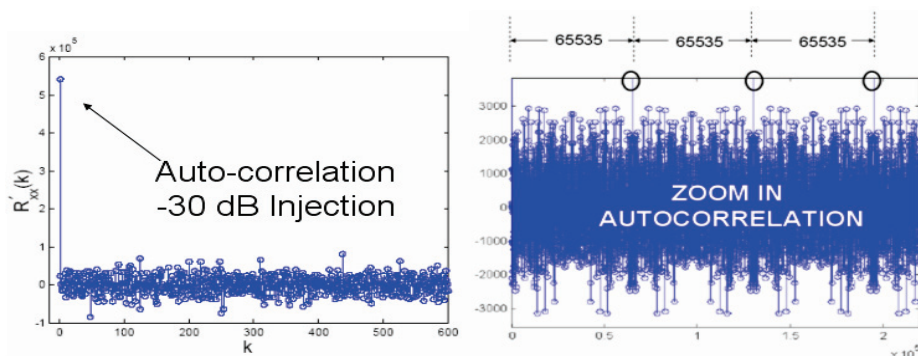


Figure 12.6 Auto-correlation of truncated Kasami sequence.

bit Network Identifier and 12-bit Transmitter Identifier pair would be used to pre-load the Kasami code sequence generator in the transmitter's RF Watermark generator and to produce the Reference Sequence in the RF Watermark receiver. The large auto-correlation peak on the left graph gives a positive identification of a transmitter in a DTxN. The autocorrelation peaks are clearly visible in the zoomed-in view over most of the length of an 8-VSB field on the right, in spite of the truncation. Note the peaks that appear every 65,535 symbols, at the ends of the first 3 complete sequence periods. The auto-correlation peaks (circled) extend off scale in this zoomed-in view.

The largest source of interference to the RF Watermark identification process comes from the in-band 8-VSB host signal, especially when using low injection levels for the RF Watermark. Figure 12.7 is a simulation showing a strong main signal and two naturally occurring echoes at -5 dB and -10 dB, relative to the main signal. When these signals arrive at an RF Watermark receiver, they all will have identical RF Watermark signatures. The graph on the left shows the auto-correlation of the RF Watermark, buried 30 dB below the host 8-VSB signal and measured during the time interval of one 8-VSB data field. Post processing of the signal using time domain averaging over multiple fields can result in an increase in the recovered dynamic range. The

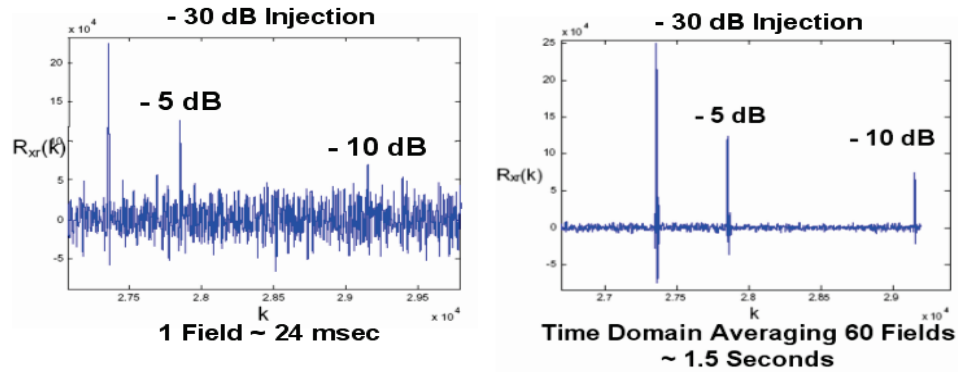


Figure 12.7 Post processing to increase dynamic range.

measurement time of 60 fields, or approximately 1.5 seconds, shown in the right-hand plot demonstrates an 18 dB increase in dynamic range.

As can be seen from a comparison of the two plots in Figure 12.7, a test receiver could trade measurement speed for dynamic range. The optimum trade-off in a particular situation will depend upon receiving conditions. As noted earlier, an operating bury ratio of -30 dB for the RF Watermark is recommended. It also has been observed previously that consumer DTV receiver adaptive equalizers become active when echoes are on the order of 20 dB or less below the main signal. This means that it is desirable to be able to measure the effects of transmitters within a network the signals from which are at least 20 dB below the strongest signal received at any given location. The combination of these factors means that it should be possible to recover an RF Watermark when it is at least 50 dB below ($-30 - 20$) the level of the strongest host 8-VSB signal at the location where the measurement is to be made. Using post processing, RF Watermark signals arriving from multiple transmitters in a DTxN can be recovered more reliably. Using special reception techniques and search algorithms test receivers also may be able to detect the RF Watermark signatures of co-channel DTV stations, whether DTxNs or single transmitter operations, allowing identification of external interferers to DTxNs or to single transmitter stations. This potential application is discussed later, in the section on permanent field monitoring points.

An RF Watermark test receiver requires a Kasami code sequence generator configured identically to that in the transmitters and needs to know the pre-load values used to set the initial states of the shift registers in the transmitters from which it is desired to receive the RF Watermark. This advance information is necessary because over 16 million unique sequences can be generated, depending upon the pre-load, or seed, value used, and searching through them all normally would be too inefficient a process. It also can be useful for the test receiver to have information on the various time settings to which the several transmitters in a network are set. The Kasami code sequence generator configuration is established by ATSC A/110 [3], and the pre-load information can be obtained from the appropriate fields of the Distributed Transmission Packets (DTxPs), which are transmitted specifically to make that information available to test equipment. So, too, can the transmitter timing settings be obtained from the DTxPs. Thus, an understanding of the syntax and semantics of the DTxP, as it relates to maintenance, verification, and optimization of a DTxN, can be quite useful.

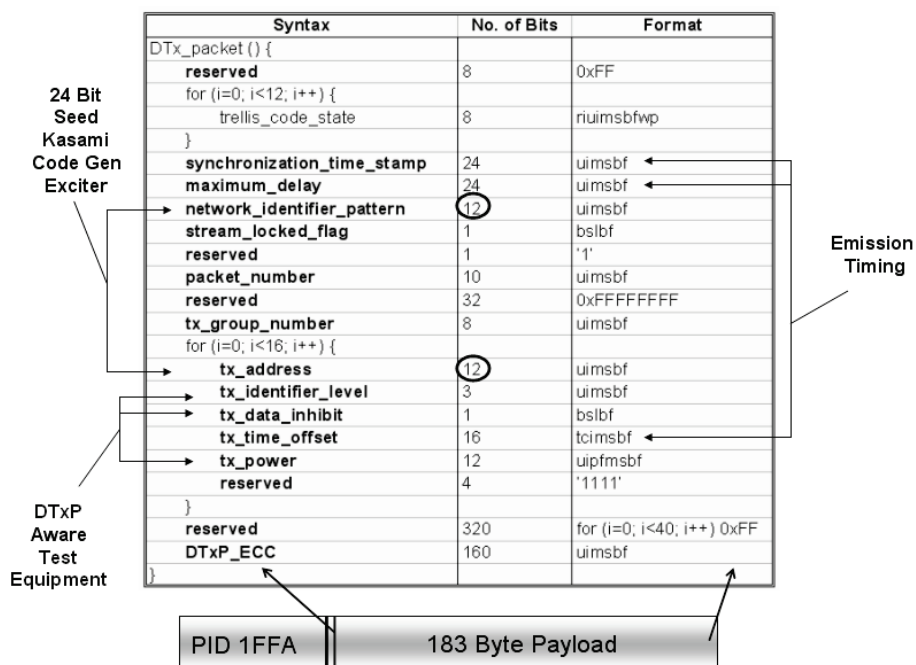


Figure 12.8 Fields in DTxP important for maintenance of DTxN.

Distributed Transmission Packets (DTxPs) are used for the operational control of slave exciters in Distributed Transmitters (DTxTs) and Distributed Translators (DTxRs). Figure 12.8 shows the structure of a DTxP, which always has a PID of 0x1FFA, indicating an Operations and Maintenance Packet (OMP). An OMP begins with an OM Type field that indicates its purpose, followed by a payload of 183 bytes. For DTxPs, the OM Type value ranges from 0x00 to 0x31, with 0x00 used for DTxTs and values above that used for tiers of DTxRs.

DTxPs are inserted into the transport stream by a Distributed Transmission Adapter (DTxA), usually located at the studio, for distribution to and control of each DTxT or DTxR, over an STL or over the air, respectively. The DTxPs are then broadcast over the air, where they may be used by field test equipment to read network parameters. By collecting information from the DTxPs, monitoring equipment can display the data and analyze the network.

A DTxP passes through three stages of formatting as it moves from the service multiplexer, through the DTxA to the transmitters, and then through the transmitters. At each stage, the DTxP has a different set of semantics, as explained in ATSC A/110 [3]. When the DTxP is demodulated by a DTx Network test instrument, the information that it carries could be very useful. For example the instrument instantly could discover the data in the 12-bit Network Identifier and 12-bit Tx Identifier fields and could learn the number of DTx transmitters using the RF Watermark and currently on the air (`tx_power` > 0). If the instrument were GPS-enabled and contained the database of assigned Identifiers, the instrument would instantly know its geographic location. Then, from a database lookup of the locations of all operating DTx transmitters, a map easily could be displayed. The instrument could display the current operating ERP (`tx_power`) and the distance and direction to each transmitter. This all could happen within seconds after receiving the first DTxP.

12.1.2 Application to Distributed Transmitters and Translators

The RF Watermark can be used effectively for the identification and measurement of both distributed transmitters (DTxTs) and distributed translators (DTxRs or EDTxRs).

In addition, by phase inverting all of the bits of the RF Watermark sequence during a full field of 259,584 symbols, an optional, low data rate, independent Return Channel can be created. The possible uses of a return channel are discussed in the section below on establishing a monitoring point.

12.1.3 Application to On-Channel Repeaters

In the RF and IF amplifier types of DOCRs, there is no opportunity to substitute RF Watermark signals, and the same RF Watermark received from the main transmitter simply must be repeated. This means that the RF energy emitted from these types of DOCRs isn't uniquely identifiable.

The EDOCR equalizes and regenerates the baseband signal and therefore also can erase the RF Watermark carried by the main DTV transmitter signal it has just received. A unique Kasami sequence assigned to the EDOCR then can be generated and inserted on the 8-VSB host signal.

The optional return channel could also be established with an EDOCR through modulation of the RF Watermark. If the design of the EDOCR is aware of DTxPs, some basic control over the EDOCR also could be achieved, but not timing adjustments.

This means that an EDOCR can be treated in some ways as a distributed transmitter since it can transmit relatively higher power (in comparison to RF and IF DOCRs) and can have the option of inserting its own RF Watermark in the output signal.

12.1.4 Measurement of the Channel Impulse Response

One of the most important measurements that the RF Watermark enables is the Enhanced Channel Impulse Response (ECIR).

The graph on the left in Figure 12.9 represents the result of a conventional Channel Impulse Response (CIR) calculation. This measurement takes data samples in the frequency domain and transforms them to the time domain by performing an Inverse Fast Fourier Transform (IFFT) on the data samples. The result of this transformation is the impulse response of the transmission channel. The x -axis is shown as time in this simple example but also could be scaled to represent distance, in units of miles or kilometers. The y -axis, in dB, represents the signal strengths of all arriving signals that have taken various propagation paths from the transmitting to the receiving antenna. The received signal that is the strongest is normalized to zero dB on the y -axis and to zero time on the x -axis. The main (strongest) signal is treated as the reference and would be the signal traveling the direct, line-of-sight path in a Ricean channel.

The normal CIR shown on the left represents a classical Ricean channel, with a direct line of sight path represented by the strongest signal and shortest delay. A series of trailing echoes is shown; these would result from reflections and diffractions from natural and man made objects in and near the direct propagation path.

The conventional CIR can be combined with the power of the auto-correlation property of the RF Watermark to create substantial synergy. This combination enables the in-service measurement, verification, and optimization of a DTxN. The new measurement technique is termed the Enhanced Channel Impulse Response or ECIR.

Through signal processing methods outside the scope of this RP, the Kasami sequence can be recovered from the host 8-VSB signal on which it is carried on each delayed path. Once it is

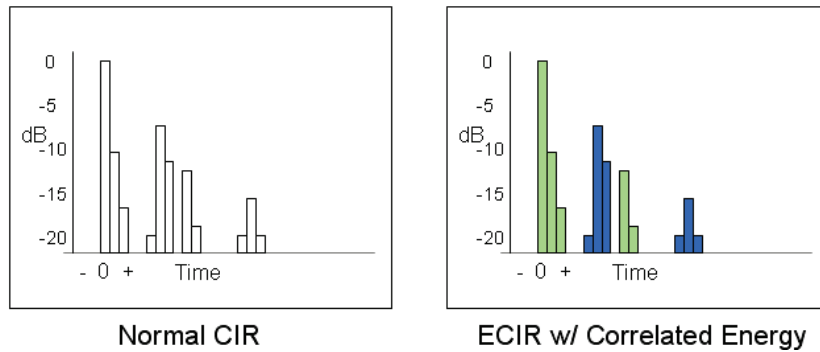


Figure 12.9 Transmission channel impulse response.

separated from its 8-VSB host, auto-correlation can be performed on the recovered Kasami sequence to identify the specific DTxT that created the RF energy received over a particular delayed path. This process would be applied to the signals from all paths detected in the CIR that are within the resolution capability of the technique. Given the coding gain of the selected Kasami code sequence, combined with averaging over a sufficient number of repetitions of the code, a minimum dynamic range of ~ 50 dB can be expected with a resulting minimum signal-to-noise (S/N) ratio of 10 dB. Using a bury ratio of 30 dB for the RF Watermark emitted by all DTxTs, this means that an RF Watermark signal can be detected when its host 8-VSB signal is at least 20 dB down from the main, or strongest, signal.

The 20 dB value is important because, in a DTxN, advantage is taken of the fact that receiver adaptive equalizers only start operating at echo signal levels greater than 5–6 dB below the reception threshold of approximately 15 dB S/N. They treat lower level echo signals (and signals from alternate transmitters), even though coherent with the main signals, in the same way as other noise in the channel.

Returning to Figure 12.9, the left hand graph shows the successful identification by the CIR of all paths having signal levels greater than -20 dB relative to the main signal. The right graph shows the enhancement that results from the synergy of the ECIR measurement. The normal CIR displayed on the left is enhanced on the right with knowledge of the sources of the correlated energy received. Assignment of a unique identifying color to the display of all of the received energy correlated to a given Kasami sequence allows immediate recognition of the sources of the various impulses in the channel impulse response presentation. Thus, it can be seen in Figure 12.9 that two transmitters are contributing to the received signal, that their signals are displaced in time at the receiver, and what the amount of the time displacement is. Moreover, it is readily apparent which lower level echo patterns are originating from which transmitter.

The largest green and blue energy peaks in Figure 12.9 are termed “SFN signals.” SFN signals are the signals that a reception point receives directly from each of the multiple transmitters in a single frequency network. All the other green and blue energy peaks are considered “natural echoes.” Natural echoes are static and dynamic signals taking multiple paths from transmitter to receiver that are created by diffraction and reflection of signals from nearby (stationary and moving) objects. In addition to such graphic presentations of the ECIR, information gleaned from the RF Watermark and the DTxP can provide numeric values for such parameters as transmitter power levels, received signal levels, transmitter delay settings, total delay spread at the receiver, $C/(I+N)$ between various combinations of transmitters, and the like.

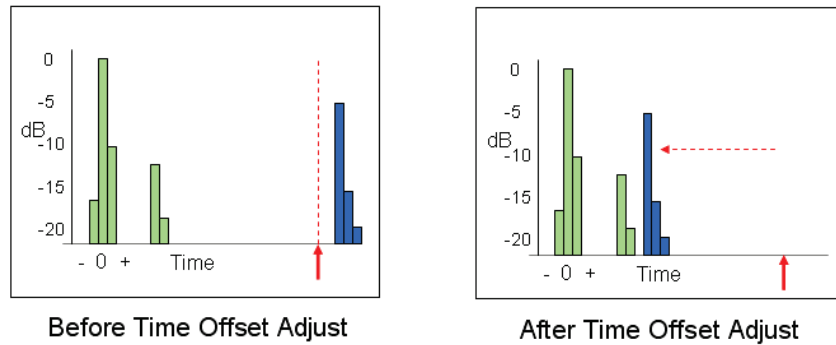


Figure 12.10 Timing adjustment (post-echo).

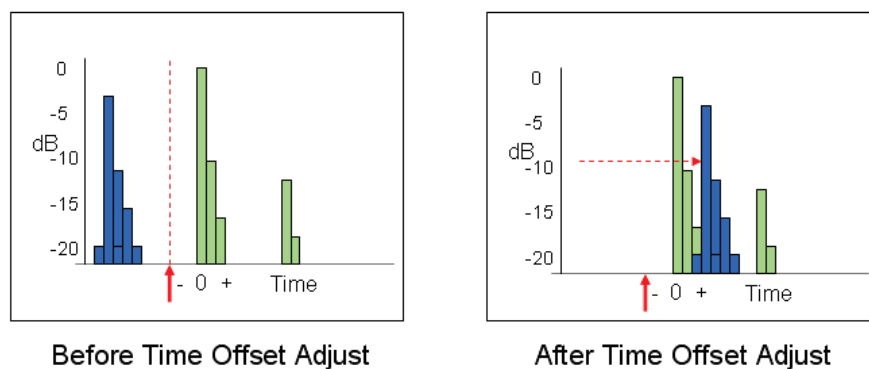


Figure 12.11 Timing adjustment (pre-echo).

12.1.5 Measuring and Adjusting Transmitter Emission Timing and Power

Figure 12.10 shows on the left graph an observation of a strong post echo. The red marker is intended to indicate the maximum acceptable delay, a design parameter for the particular DTx transmitter at the specific monitoring location. The right graph shows the result of a timing adjustment entered as a change in the value of the `tx_time_offset` field directly addressed to the particular transmitter and carried by the DTxP. When the value change in the DTxP was accepted by the slave DTx exciter and executed as a command, all of the energy from the DTxT represented in blue would make a step transition to the left, indicating an advance in its relative emission time.

The new value for the `tx_time_offset` field would be chosen by the network designer to correspond to the exact advance in timing needed to shift the received signal to the optimum time offset for the measurement location. As the echo handling performance of the population of 8-VSB receivers improves or as knowledge is gained concerning operation of the particular network, further optimization of timing parameters easily can be performed.

A pre-echo is the energy that arrives at a receiver before the main signal, as shown in blue on the left graph in Figure 12.11. In this example, the strong pre-echo energy is from a transmitter belonging to the DTxN.

The pre-echo handling capabilities of legacy receivers is known to be quite limited, with many capable of recovering data from signals having just a very few microseconds of pre-echo. Control of the pre-echo energy of SFN signals in critical field locations is one the most important design

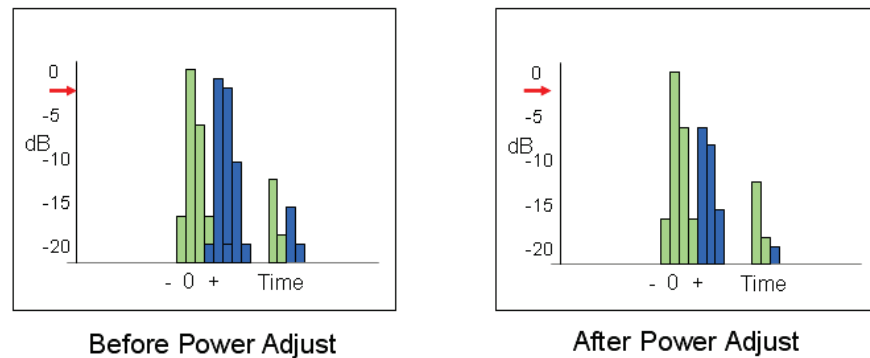


Figure 12.12 Power adjustment.

constraints that early DTxN designers will face. This constraint can be somewhat relaxed with current receivers and is expected to be relaxed even further with future receiver designs.

The right graph in Figure 12.11 shows the result of a timing adjustment entered as a change in the value of the `tx_time_offset` field, directly addressed to the transmitter represented in blue and carried to it by the DTxP. The pre-echo shown on the left graph is delayed and this causes a shift to the right, as shown on the right graph. The energy from the transmitter represented in blue is now a post-echo, and the network design dictates its optimum temporal position relative to the main signal.

The previous examples show that the delay spread of SFN signals is an important parameter in network design. When delayed signals from alternate transmitters exceed a certain time offset and are stronger than -20 dB relative to the main signal at critical locations, they must be taken into account. They are the source of network internal interference. Delayed signals weaker than -20 dB relative to the main signal will appear as non-coherent to a consumer receiver and will be treated as noise.

The `tx_power` field in the DTxP allows for adjustment of the output powers of transmitters in a DTxN. Figure 12.12 shows the received signal power of the transmitter depicted in blue before and after application of a command in the DTxP to decrease transmitter power output.

12.1.6 Identifying Interfering Transmitters

In analog television broadcasting, the picture tube serves as a window into the RF world, and the existence of co-channel interference can be discovered visually, usually by observing the call sign of a faint image superimposed on the desired signal by an interfering station. With DTV transmission, things are quite different. The Forward Error Correction (FEC) used corrects many errors that would otherwise occur in the terrestrial channel, which is its purpose. Receivers have a finite amount of error correction capability, and errors will be dammed up (corrected) behind the FEC and not be visible on a DTV display device.

When the FEC capability becomes overrun by increasing errors, first a trickle of errors starts appearing on the display; then, a flood of errors quickly follows. No pre-warning of a pending interference condition can be gleaned simply by looking at a display, as is the case with analog signals. Digital receivers react to co-channel DTV interference of a particular level relative to the desired signal, producing errors in the same way as they would in reaction to noise of the same relative level. Thus, just as there is little warning of impending failure (due to the “cliff effect”)

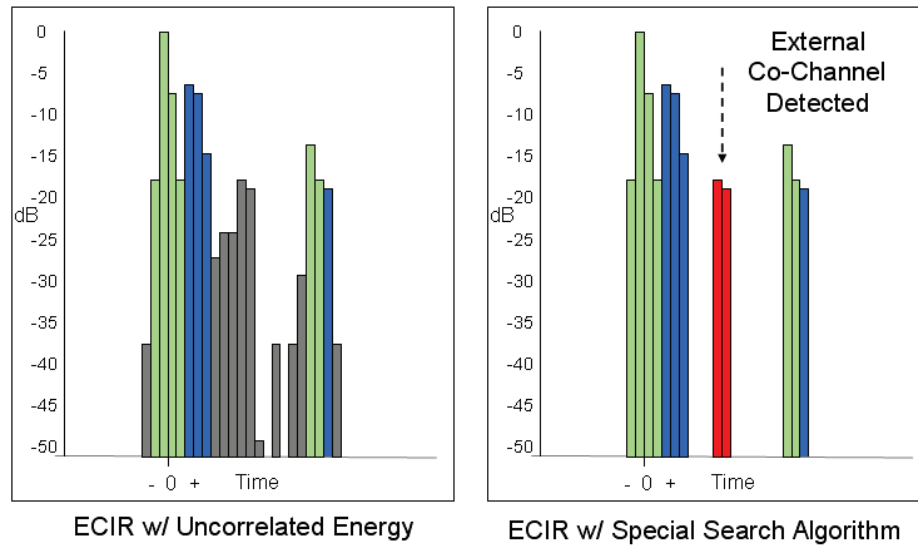


Figure 12.13 Discovery of co-channel interference.

when the received signal level drops close to the noise level, there similarly is little warning as an interfering co-channel signal approaches the level at which the FEC is no longer able to correct errors and reception fails. Because there can be no visual indication of the source of co-channel interference, often the only way to confirm the source is to turn off one or both of the transmitters — a test solution that is not desirable for either station, that can be logistically difficult, and that would normally be arranged only after real interference already has occurred. The RF Watermark can allow discovery of the source of co-channel interference without turning off either transmitter — equivalent to non-destructive testing. Depending upon when they are applied, RF Watermark measurements can indicate the presence of co-channel interference before it becomes a problem.

In Figure 12.3, the left graph shows an increased display range in the y-axis, down to -50 db. The energy shown in grey represents uncorrelated energy. Such energy cannot be auto-correlated with the RF Watermark of the DTx network, but is present in the CIR. When additional post processing such as time domain averaging over a larger number of DTV fields is used to increase the dynamic range, an attempt to auto-correlate on the known tx_address, as carried in the DTxP, can be tried again with the increased sensitivity. In the example shown in the left graph, such an attempt was not successful in correlating any new energy in the aggregate received signal.

In such a situation, a list of candidate DTV co-channel interfering stations could be loaded from a database into the RF Watermark receiver. The Kasami sequences representing the candidate stations then would be scanned. If a positive auto-correlation result occurred with any of these Kasami sequences, it would indicate that co-channel interference was detected at some measurable level. The database then would provide the identity of the interfering DTV station.

The graph on the right shows the result of such a successful identification of co-channel interference. Only correlated energy is displayed in this graph, with the energy in red representing the DTV interfering station external to the network. Whenever there is uncorrelated energy less than -20 dB, it is a good time to check for possibly identifiable external DTV interference. For this method to work, all stations involved must be transmitting an RF Watermark. An instrument

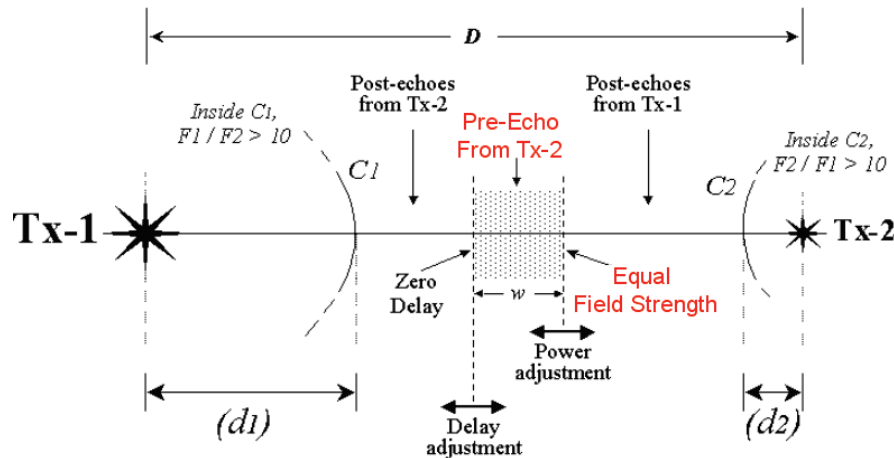


Figure 12.14a Signal overlap between DTxTs or DTxRs.

with ECIR capabilities also could be used in coverage areas of neighboring non- DTxN DTV or even NTSC stations to detect the level of co-channel signals emanating from an operating DTxN.

12.2 Selecting Critical Locations

Critical locations for RF Watermark measurements can be geographical demarcation points that the designer selects to verify network design parameters. They can be used to provide a sanity check on areas of high population density. They can be in areas with a high probability of stressing legacy receivers, such as those areas predicted to have large pre-echoes or nearly equal signal levels (i.e., 0 dB echoes). They can also be in areas in a large cell design that have the likelihood of producing long post-echo energy, outside the mask of receiver delay spread targeted by the design.

12.2.1 Locations of Signal Overlap

The locations selected should have overlapping SFN signals from two or more transmitters. The area of overlapping SFN signals should have less than a 20 dB difference in levels between the signals from the several transmitters.

Figure 12.14a shows an example of the overlap of SFN signals between two DTxTs or DTxRs. Two areas shown could be of concern for legacy receivers. They are the grey area in the center, which has a large pre-echo from Tx-2, and the location indicated as having equal field strength (0 dB echo). These areas that can be stressful on consumer DTV receivers will be used as examples of places to be monitored when network adjustments are made.

Figure 12.14b shows a zone that can be challenging for service from a DOCR. There always will be an internal delay through the DOCR that adds to the transit time over the paths to and from the repeater. The signal from the main transmitter will always arrive first. In areas where the main transmitter signal is stronger, the DOCR signal becomes a long post-echo. In areas where the DOCR signal is stronger, the main transmitter signal still always arrives first and therefore acts as a pre-echo. To mitigate this effect, DOCRs should be used where there is terrain shielding to prevent long, strong pre-echoes, which can be problematic. An RF or IF processing DOCR also is not capable of inserting an RF Watermark. Direct identification of the RF energy emitted by such

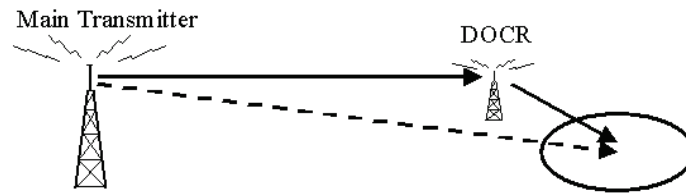


Figure 12.14b Signal overlap area of DOCR or EDOCR.

a DOCR therefore is not possible, and the measurement techniques described above cannot help in configuring systems using such devices.

An EDOCR, on the other hand, potentially can provide a much cleaner signal than an RF or IF processing DOCR, but it has an even longer processing time — on the order of 5 microseconds. The EDOCR, however, because of its signal processing can insert a unique RF Watermark. While substantial terrain shielding is necessary for use of an EDOCR, the possibility to include an RF Watermark in its output means that the measurement techniques described can be applied to the implementation of systems employing them.

12.3 Network Adjustment

When conditions exist that could be stressful to legacy receivers, network adjustments often can be used to mitigate them,

12.3.1 Setting Offsets to Mitigate Interference

Referring again to Figure 12.14a; it will be used as a starting point for the following discussion of network adjustments. The figure shows a possible area of concern for legacy receivers, the large gray “Pre-echo from Tx-2” area in the center of the diagram, bounded by the “zero delay” line on the left and the “equal field strength” line on the right. For purposes of this discussion, the gray center region is assumed to have a high population density.

Figure 12.15 shows the results of an attempt to mitigate the situation as much as possible. There is assumed to be a low population density area, to the right of the gray area of Figure 12.14a, toward Tx-2, of which advantage is taken in this example. First the power of Tx-1 is increased. This moves the equal field strength line to the right, into the center of the target area of low population density. This causes the pre-echo from Tx-2 area to become temporarily larger. The larger area of pre-echo from Tx-2 then is reduced in size and moved into close proximity to the equal signal strength boundary by increasing the delay of Tx-2. This places both the narrowed Pre-Echo from Tx-2 area and the equal field strength line in the center of the target region of low population density. The conditions of pre-echo and 0 dB echo, known to cause difficulty for reception with legacy receivers, have been shifted from the area having the larger population density into one where there will be less impact. Furthermore, by placing the two conditions in proximity to one another, the stress on receivers is reduced through the virtual elimination of strong pre-echoes.

To confirm the intended operation of the network, an instrument with ECIR capability could be taken to the center of the target low population density area. The values for power and delay adjustments could be calculated and sent to Tx-1 and Tx-2 via the DTxPs. The ECIR then would be used to verify that approximately equal signal levels were received from Tx-1 and Tx-2 and that approximately equal signal arrival times existed at the measurement point. The overlapped

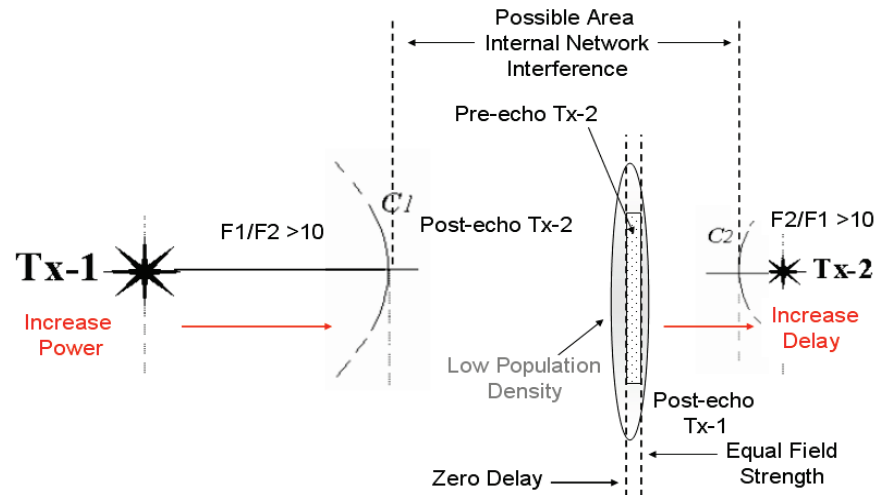


Figure 12.15 Offsets to mitigate interference.

coverage area having the higher population density now would receive only post-echo energy, which should be less problematic for legacy receivers.

12.4 Field Verification

Field verification can be used to validate a network design, confirming adjustments made to optimize the network. This verification would be carried out at geographic locations selected to be optimum for making such field observations.

Two types of DTxN test equipment are envisioned for making field observations and are described here for the purpose of showing the resources for network monitoring and adjustment built into the A/110 transmitter synchronization standard:

- *Portable hybrid*: a spectrum analyzer for signal capture, networked to a notebook PC executing the needed signal processing and analysis software
- *Purpose-built*: a single-piece test instrument designed for application to DTxN measurements that would use GPS data and be DTxP aware.

The principal differences between the two measurement concepts are that the portable hybrid equipment would be operated manually, while the purpose-built instrument would use information from the received DTxPs to automate its processing. Along with a built in database of all DTxNs and neighboring DTV stations, the purpose-built instrument could use the GPS data to provide mapping and timing capabilities to assist the engineer making field measurements. These two DTxN field test equipment concepts are illustrated in Figure 12.16.

12.4.1 Field Measurement Techniques at Selected Locations

Measurement techniques using the two types of instrumentation outlined above will be described to help in understanding the processes of network confirmation and monitoring. For simplicity, omni-directional receiving antennas are assumed for these discussions. The parameter values expected to be observed at each location would be known in advance and supplied by the network designer. The fundamental purpose of the measurements described is verification that the

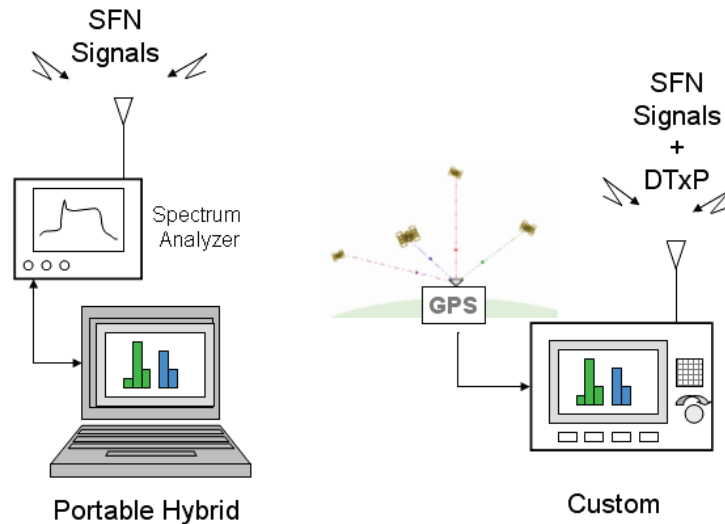


Figure 12.16 Types of DTxN field test equipment.

expected parameter values are indeed observed. The measurement process with each type of instrumentation will be described.

Starting with the portable hybrid solution, it would be set up at the selected test point. The Network Identifier and Transmitter Identifier portions of the RF Watermark code sequence preset data would be entered for each DTxT that is operational and expected to be observable at the measurement point, along with other pertinent parameters, into the software running on the notebook PC. The software running on the notebook would then control the spectrum analyzer. The spectrum of the channel shared by the DTxN then would be displayed. An RF signal data capture would occur, be digitized, and be quickly passed over the network connection to the notebook. The RF spectrum, the normal CIR, and the correlated ECIR displays would be computed and presented for operator analysis. The operator then would compare these metrics to the predicted values as a check or verification of the design parameters used.

If the purpose-built DTxN test equipment were used, advantage could be taken of an embedded GPS receiver providing a precise 10 MHz frequency reference and the 1PPS clock tick used for internal timing by the DTxN. An internal 8-VSB receiver could parse the transport stream and decode the payloads of all received DTxPs. This functionality could enable such features as a geographical map that could display the location of the measurement point and the locations of all DTxTs of interest, display the parameter values to which the various transmitters were set, and determine whether the measured signals matched the predictions for them at the measurement location and what any differences were. The purpose-built instrument also could have pre-loaded databases containing all field locations to be used for network verification, through which it could confirm that it is situated at a measurement point, and all locations of the transmitters in other networks or even single-transmitter operations, through which it could identify co-channel interference sources.

From a network operations control point, a command could be given to a particular DTxT by sending new parameters in a DTxP to adjust the transmitter's delay offset or power level. Since the purpose-built instrument would be reading the DTxPs and would see instantly any change in parameters sent to a given DTxT, a new analysis using the ECIR measurement could be made

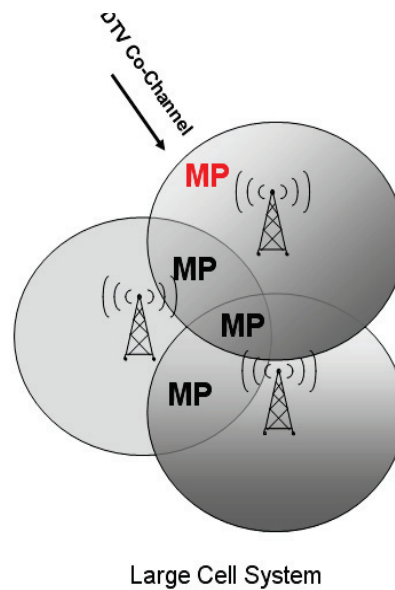


Figure 13.1 Critical measurement locations for large cell design.

automatically. The results obtained before and after the change could be presented in a matter of seconds with no further operator intervention. These results could be stored in the database and used as a benchmark against which future measurements could be compared.

Both types of equipment shown in Figure 12.16 are capable of making accurate field measurements in a DTxN. The intelligence of a purpose-built instrument should make it easier to use, may allow more repeatable results, and could provide a wider range of features in a more compact configuration than might be possible with the hybrid instrument approach.

13 PERFORMANCE MONITORING OF DISTRIBUTED TRANSMISSION NETWORKS

Monitoring of a DTxN starts with field verification prior to network operation and continues with operational monitoring once the DTxN is on the air and delivering services. Both the need for and benefits of monitoring are discussed in this section. The useful information to be gained from short term monitoring is considered along with the value of establishing longer term or permanent monitoring points.

13.1 Short Term Measurements and Monitoring

The primary considerations for short term monitoring include field verification and multiple measurements at critical locations.

13.1.1 Use of Field Verification

Field verification and measurement techniques at critical locations were discussed in Section 12. Some of the benefits of field verification are:

- Quick detection of errors in network planning or setup
- Objective validation of the network design
- Interference analysis, both internal and external to the network

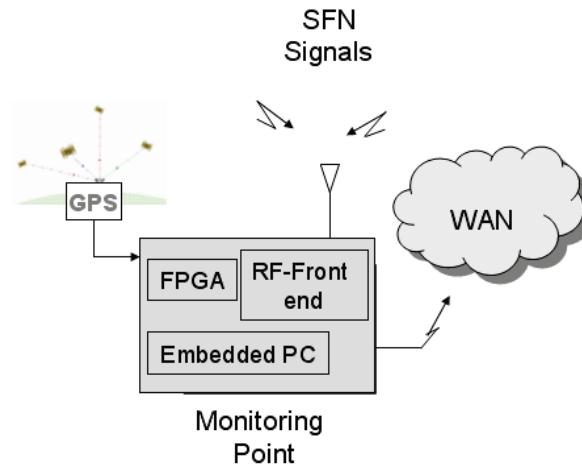


Figure 13.2 Example field monitoring point equipment.

- Obtaining of historical (archive) data from critical locations as a baseline
- Collection of empirical data to improve design tools

13.1.2 Multiple Measurements at Critical Locations

The example network in Figure 13.1 shows that the critical areas for measurements are in the coverage overlap areas associated with the three DTxTs used in this large cell design. Network internal interference present will be concentrated in these areas. Monitoring Points (MPs) for network internal interference that could be placed in the overlap areas are shown in black. The red (MP) could be used to observe and benchmark possible DTV co-channel interference at the location(s) expected to have the highest susceptibility to such interference. Any of the critical locations would make good sites for DTxN field monitoring points. For maximum efficiency in the monitoring system, the single monitoring point in the area where all three signals overlap would be used to monitor the entire network.

The example in Figure 13.1 shows a Large Cell design. The transmitters would be spaced by longer distances than in either Small or Micro Cell designs. The longer the distance between a transmitter and a receiver, the higher the probability that variations in signal strength or fading will occur over a particular propagation path. To characterize the typical path fading, a monitoring point can be established to observe diurnal and seasonal fading. Empirical data collected by such a monitoring point can help determine the real world fade margin needed and validate the selected DTxN optimization parameters. The efficacy of any mitigation efforts undertaken to benefit legacy receivers in certain areas could also be confirmed.

Another benefit of monitoring would be to discover potential co-channel interference from neighboring DTV stations as the RF spectrum is repacked as part of the transition to DTV. Such co-channel detection and measurement could be a very helpful tool for broadcasters in early detection of the presence of external interference into a DTxN. Prerequisite is that co-channel DTV signals carry an RF Watermark to be detected by this means.

Figure 13.2 shows a basic configuration for the equipment that might be used in a remotely monitored short-term or long-term monitoring system. Such equipment would comprise a high quality RF front-end providing an IF output in the selected frequency range, which would be sampled by a 12-to-14 bit A/D converter to create a digital IF output. An FPGA would process the

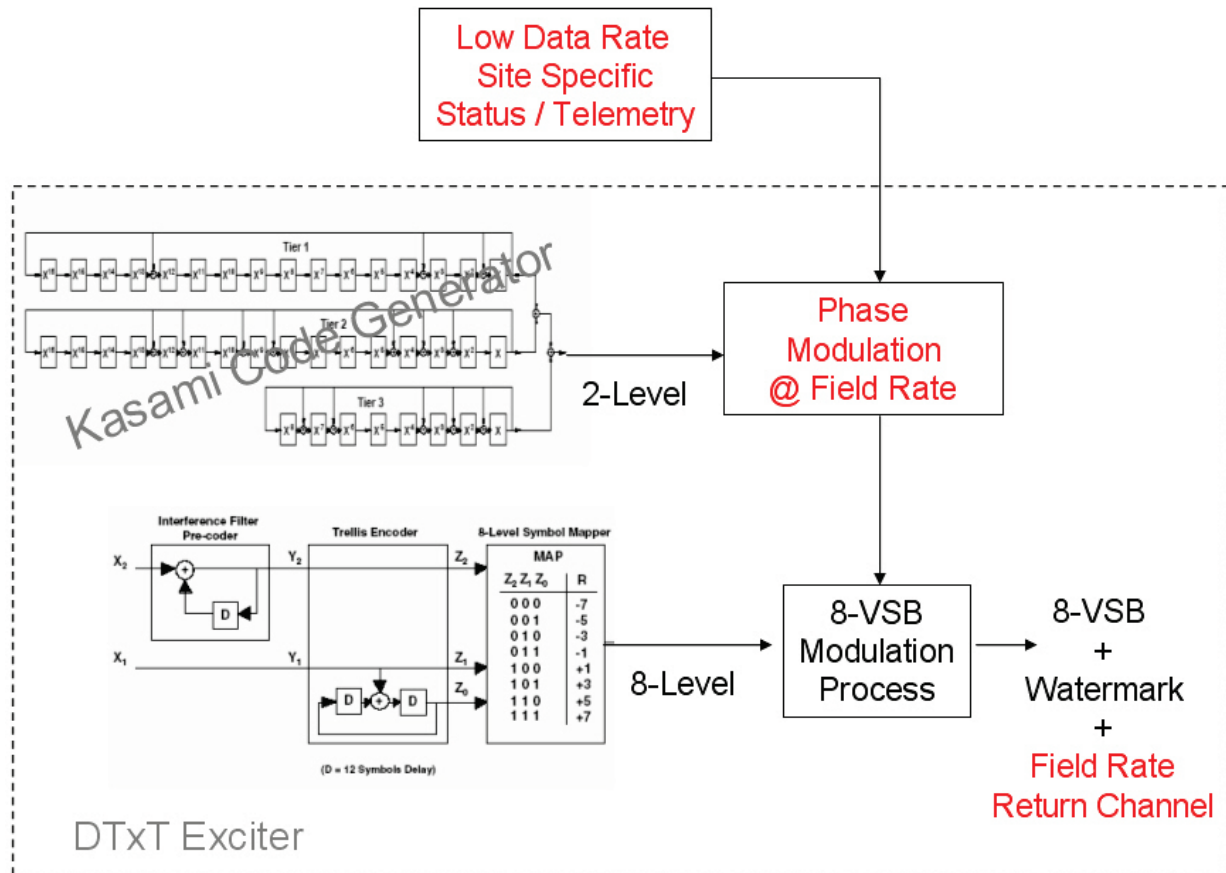


Figure 13.3 Site return channel.

digital IF and pass the sample data to an embedded PC for analysis and remote communications. Only the measurement results would need to be communicated, so a low-rate data channel could be used for the data return path.

13.2 Long Term Operational Applications

While the original purpose of the RF Watermark was for measurements, its ability to carry data at low rates provides opportunities for applications in the operation of the network when monitoring points are permanently installed. Such monitoring points would simultaneously collect data on network performance through measurements and on network operation through telemetry sent back by the respective DTxTs.

13.2.1 Establishing Monitoring Points

The basic theory of the RF Watermark was covered in Section 21, where a technique was described to create a Return Channel (RC), effectively allowing each DTxT to transmit an independent data signal on the broadcast channel using a CDMA technique. While there are no restrictions on the data that can be carried on the RC, given the relatively low data rate (the 8-VSB data field rate, or slightly over 40 b/s) carriage of telemetry from each transmitter to a central monitoring point is a potential application. Because the RF Watermark is designed to allow

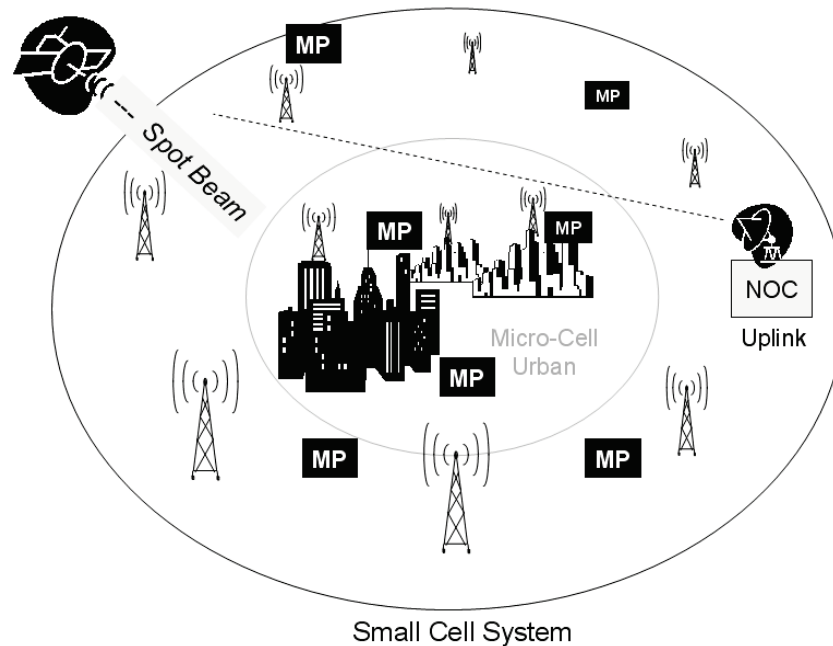


Figure 13.4 Hybrid DTxN design.

reception from (and identification of) several transmitters operating on the same broadcast channel at the same time, data from each of the transmitters can be transmitted to such a monitoring point at the same time with no interference between the multiple data channels.

Figure 13.3 shows the block diagram of a circuit for modulating the output of the Kasami code sequence generator before its code sequence modulates the 8-VSB symbols to become an RF Watermark. Modulation of the RF Watermark signal is by phase inversion of the code sequence associated with each transmitter. Phase inversions occur on data field boundaries, as described in Section 12.3 of ATSC A/110 [3]. The on or off state of the return channel data modulation is signaled by a 1-bit field (`tx_data_inhibit`) in the DTxP data addressed to each transmitter.

13.2.2 Permanent Installations

Because much the same equipment is required to receive the RF Watermark for measurement purposes and for reception of RC data, return channels may be of interest to broadcasters operating systems employing small-cell and micro-cell transmitters at low power to cover urban canyons, for creating signal level hot spots, and the like. The same monitoring points both would conduct measurements of network performance and would serve as collection points for status and telemetry data from the network DTxTs.

Figure 13.4 shows an example hybrid DTx Network design in which the broadcaster has chosen to combine a Small Cell approach to maximizing the DTV service area with a Micro-Cell design to fill in the urban areas and city canyons most effectively. Some of the DTxTs are likely to be small, low power devices located on the tops or sides of buildings, and the cost of control circuits to these places would not be commensurate with their value. The logistics of such sites points to use of the RC for status and telemetry from these DTxTs. With the DTxP delivering downstream control data and the RF Watermark returning the status and telemetry data, a cost-

effective remote control system can be configured for such situations. As before, MP sites would be selected so as to minimize their number and to cover the largest number of DTxTs with a single device, thereby maximizing the efficiency of the monitoring network both for measurement purposes and RC communications.

Carrying the hybrid concept one step further, the DTV transport stream distribution to all DTxTs in Figure 13.4 is also a hybrid design. Some DTxTs in critical areas receive their data inputs via a microwave or fiber STL. For the others and for redundancy to DTxTs in critical areas, a spot beam from a satellite with a footprint covering the station's service area provides the data stream feed. The small physical sizes and the locations of low power DTxTs in micro-cell applications allow for easy deployments, and the Return Channel can be used synergistically to close the loop for remote control.

14 IMPLEMENTATION MATTERS

The successful planning of a multiple transmitter network requires attention to a number of important elements addressed in previous sections of this document. Here, specific implementation issues are considered relating to both Distributed Transmission (DTx) and Baseband Equalization Digital On-Channel Repeater (EDOCR) systems.

14.1 Distributed Transmission

Implementation considerations relating to a distributed transmission network include: 1) STL and over-the-air path delay variations, 2) RF watermarking, and 3) the optional GPS-locked operating mode.

14.1.1 Consideration of STL and Over-the-Air Path Delay Variation

There are several steps in the basic method of setting a slave transmitter's delay. First, when each DTxP is received at the slave, the transport delay is measured by comparing the Synchronization Time Stamp (STS) value to the local GPS clock's 1 pulse per second (1 PPS) reference. Then, by reading the maximum delay and offset delay values from the DTxP and subtracting the transmitter delay value, the additional time delay required in the modulator can be determined. This method assumes that the time delay variation in the MPEG-2 Transport Stream distribution system (i.e., the STL) is constant. This is called the "stream locked mode" of DTx slave operation. This mode is indicated in the DTxP by setting the single bit in the stream_lock field to 1 at the DTxA.

There are several factors that can cause STL delay to vary. These include propagation effects, STL circuitry implementation, and physical path length changes that inherently occur in satellite systems due to station-keeping maneuvers. Similar effects occur in over-the-air delivery of signals to DTxRs.

In conventional, single-transmitter (non-DTx) operations, a slow change in STL time delay is of no consequence. As long as the STL-induced Doppler shift is well within consumer receiver tracking capability, the variations in timing will not affect demodulation. But in a DTx system, any STL time delay variation can directly skew timing in signal overlap areas, which, in turn, can degrade consumer receiver delay equalization capability.

In situations where STL delay is not constant, the basic timing method will result in the emission timing of slave transmitters changing by the same amount as the STL delay is changing. If the STL time delay variation changes significantly between DTxPs, then there is no way for the slave transmitter to measure the delay change, let alone to compensate for it.

When STL delay variation occurs, there is a timing mode that can be used to correct the consequent errors. This method is called GPS lock, and it is indicated by setting the `stream_lock` bit to 0 at the DTxA. When a slave transmitter receives a `stream_lock` bit set to 0, it must lock its local MPEG-2 Transport Stream clock to the GPS 10 MHz reference by the ratio 433998/223795. Because the ATSC symbol rate is locked to the MPEG-2 Transport Stream clock by the ratio 313/564, the ATSC symbol rate, in turn, will be locked to the GPS 10 MHz reference by the ratio 1539/1430.

With this method of regenerating the MPEG-2 Transport Stream clock frequency, the only limits on the rate or extent of STL time delay variations are the ability of the MPEG-2 Transport Stream clock recovery circuits to track the frequency changes and of the FIFO size to buffer the delay changes.

In GPS mode, it is recommended that the stream-locked mode initially be used to acquire and set a transmitter's timing. Once the timing is correct, and assuming that the stream lock flag is set to 0, then the transmitter must switch to GPS mode.

Another way to analyze this issue is to consider that, when an STL is changing its delay, it is changing the MPEG-2 Transport Stream frequency. (The frequency skew is proportional to the time derivative of delay/phase.) If the correct MPEG-2 Transport Stream clock frequency is restored at the slave transmitter, then the timing variation vanishes automatically. In other words, continuously correcting the timing to remove STL delay variations is equivalent to regenerating the correct MPEG-2 Transport Stream clock frequency. If you have one, you have the other. For this reason, the GPS mode PLL solves the STL delay variation problem.

In summary, if the STL system produces time delay variations of more than a few microseconds, the network should be operated in GPS-mode for MPEG-2 Transport Stream clock recovery. GPS mode is generally required for operation of DTxRs.

14.1.2 RF Watermark Implementation

When an RF Watermark is being transmitted, the equalization system of a DTxT or DTxR operating with adaptive equalization may require modification to account for the presence of the RF Watermark — particularly at the higher bury ratios such as -18 dB.

Designers of test and teasurement (T&M) equipment also may wish optionally to subtract the RF watermark's 2-VSB symbols prior to calculating SNR, EVM, and MER and prior to creating graphical displays such as constellation and eye diagrams. T&M equipment may be able to calculate these parameters and graphical displays either with or without the RF Watermark removed.

In some evaluation situations, the true performance of the transmitter will be obscured unless the RF Watermark is removed. In these cases, the T&M equipment best would be operated with the RF Watermark switched off.

In other situations, use of T&M equipment should include the effects of the RF Watermark. For example, when looking at the nonlinearity of constellation lines on a graphical display in the presence of a high amplitude RF Watermark, the RF Watermark should be included in the display.

14.1.3 DTxA Design for GPS Mode Operation

When operating a network in GPS-locked mode, there are several ways of locking the MPEG-2 Transport Stream bit rate to GPS. One way is to use a service multiplexer or encoder that includes the capability of locking its output clock rate to an external 10 MHz reference. Care must be

exercised to ensure that the locking is exactly according to the ratio 433998/223795 and not to an approximation created by a direct digital synthesis (DDS) oscillator; if the locking is not exact, then buffer underruns or overruns will eventually occur at the slave transmitters.

If the MPEG-2 Transport Stream is delivered to the DTxA already locked to GPS, then the DTxA need not perform any special operations other than setting the stream lock flag to 0.

If the service multiplexer or encoder cannot lock to GPS or if there is some intervening piece of equipment that upsets the precise frequency of the MPEG-2 Transport Stream clock, then the DTxA may need to include the ability to change the clock frequency and bit rate of the MPEG-2 Transport Stream signal. The DTxA may lock its bit rate to GPS by inserting and/or deleting null packets or, less preferably, DTx packet placeholders.

Where null packets do not exist in the data stream, they cannot be deleted to reduce the clock frequency of the MPEG-2 Transport Stream. Even if null packets are not present, DTx packet placeholders will exist at a rate controlled by the system operator. In this case, enough DTx packets should be inserted to allow the system to correct the bit rate of the MPEG-2 Transport Stream, while still allowing enough DTxPs to be transmitted.

One MPEG packet is 188 bytes or 1504 bits. If the MPEG-2 Transport Stream clock frequency is at its maximum limit of 2.8 parts per million (ppm) high, that is a frequency error of approximately 54.3 Hz. To restore the MPEG-2 Transport Stream clock to its nominal value, one packet must be dropped approximately every $1504/54.3 = 27.7$ seconds. In situations where no null packets exist in the data stream, if it is desired to transmit a DTx packet once per second, then the service multiplexer should be set to inject 28.7 packets every 27.7 seconds, or approximately 1.0361 packets per second. If the desired rate of DTxPs is R packets per second, simply add approximately 0.0361 to R to allow for packet deletion.

If a DTxA includes the ability to adjust the MPEG-2 Transport Stream bit rate by inserting or dropping DTxP placeholder packets, it should also restamp the MPEG stream's program clock reference (PCR) values.

14.2 Implementation of Baseband Equalization Digital On-Channel Repeater Systems

This section provides a hardware implementation example of a short-delay baseband Equalization Digital On-Channel Repeater (EDOCR). The EDOCR system shown in Figure 14.1 consists of three sub-systems: receiving, signal processing, and transmitting. The major concerns in the design of an EDOCR are low system delay, good output spectrum shaping to meet RF emission requirements, and the ability to operate in an environment of first adjacent channel interference. The following sections provide design guidelines for implementing an EDOCR system that can satisfy these requirements.

14.2.1 Receiving Subsystem

The receiving subsystem includes the following elements: 1) pre-selector and low-noise amplifier (LNA), 2) downconverter, 3) demodulator, and 4) synchronization system.

14.2.1.1 Pre-Selector and Low-Noise Amplifier

The structure of the pre-selector and low-noise amplifier (LNA) in an EDOCR are the same as in conventional RF and IF processing DOCRs. A Pre-selector and an LNA are used to increase the selectivity and the input RF signal level from a receiving antenna.

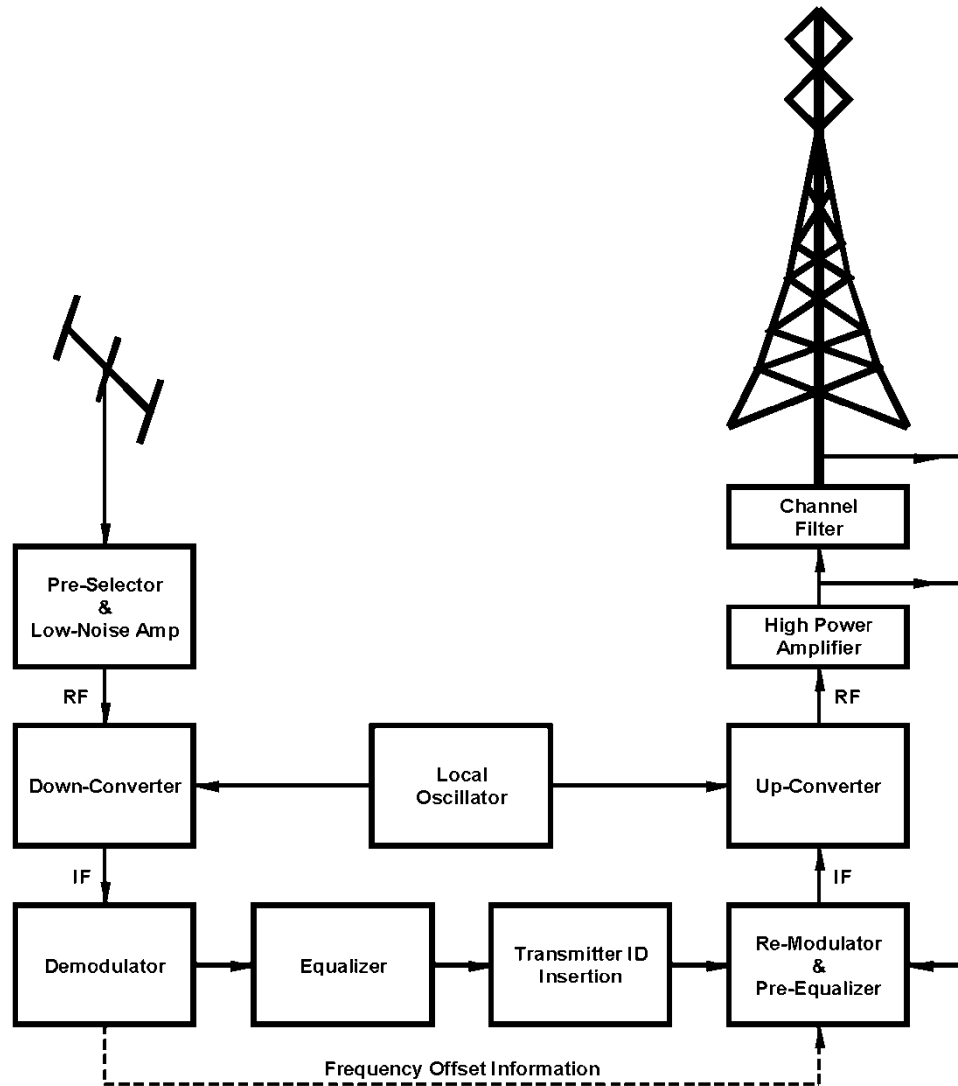


Figure 14.1 A short delay Baseband Equalization DOCR (EDOCR).

The pre-selector is a bandpass filter that eliminates the interference from image signals at $f_0 \pm (2 \cdot IF) \pm 3$ MHz frequency bands and other unwanted signals to avoid degradation of the down-converted signal. A tunable pre-selector filter can be used for easy changes of receiving channel, but a fixed pre-selector filter is also applicable. The recommended signal level of the image frequency band at the pre-selector output is at least 25 dB below the desired signal power.

An LNA is used to boost the weak received signal to an appropriate power level for down-conversion without adding much noise. The gain and noise figure of the LNA are important factors in determining the noise figure of the receiving subsystem. An LNA with a low noise figure and moderate gain is recommended. In typical applications, the LNA has a gain of 10 dB and a noise figure of 2.5 dB. The LNA should not introduce any nonlinear distortion, which can impact the EDOCR operation.

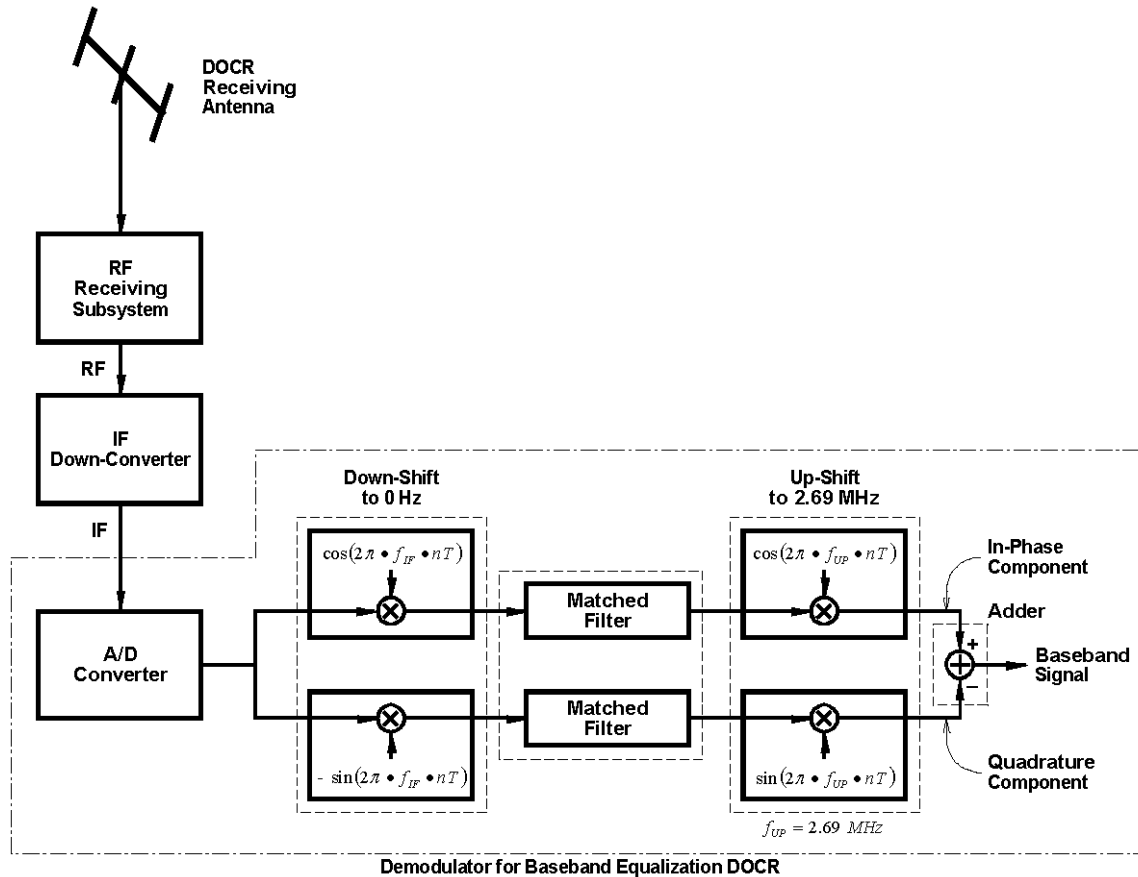


Figure 14.2 VSB demodulator for Baseband Equalization DOCR.

14.2.1.2 Downconverter

The Downconverter consists of a mixer, a band-pass filter, an IF amplifier, and an AGC control block, which are the same as in an IF conversion DOCR. The downconverter converts the desired RF signal to a fixed IF band with sufficient signal level. The frequency of the IF can be chosen by the designer. The frequency of the IF involves a trade-off. A high IF leads to better image signal rejection, whereas a low IF allows greater suppression of nearby interferers. It is also possible to use a dual-IF downconverter. In this case, each local oscillator signal should be synthesized from a common reference source. It should be pointed out that the same common reference source should be used for synthesizing the local oscillator signal for up-conversion.

The downconverter IF output signal power should be up to -10dBm . To avoid long term signal variation, automatic gain control (AGC) should be implemented in the down converter.

14.2.1.3 Demodulator

The digitized VSB signal demodulation is done in three steps: first, the IF signal center frequency is I/Q down-converted to 0 Hz; then, its I and Q components are filtered by a pair of baseband matched-filters; finally, the filtered I and Q components are up-shifted by 2.69 MHz and combined to form the baseband signal. Figure 14.2 shows the VSB demodulation system block diagram. This demodulation scheme achieves low delay because there is no need for additional

low-pass filtering to remove harmonics. The baseband matched-filter pairs are used for low-pass filtering as well as maximizing the received SNR.

A square-root-raised-cosine (SRRC) filter is generally used for matched-filtering. Theoretically, to completely remove Inter-Symbol Interference (ISI), the required number of taps of the SRRC filter is infinite. But, practically, a few hundred taps are enough to achieve a nearly ISI-free condition. As the number of taps of the SRRC filters increase, the SNR improves, but the time delay also increases. Therefore, the number of taps is determined by considering the trade-off between the SNR performance and time delay. In the EDOCR system, since the baseband equalizer with an intelligent slicer enhances the received SNR (see Section 14.2.1.4) and compensates for the distortions caused by imperfect matched-filtering, it is reasonable to reduce the number of SRRC filter taps to achieve a shorter time delay.

The recommended number of taps is between 31 and 61 when the symbols are over-sampled at 4 times the ATSC system symbol rate. The recommended number is based on performance analysis of the SRRC filter with different numbers of taps and delays. The simulation results are shown in Table 14.1. In the simulation, the number of taps of the SRRC filter in the transmitter was 521. To compare only the performance that depends on the number of taps of the matched filter, the channel was assumed to be AWGN, i.e., having no multipath distortion. From Table 14.1, for a 31-tap SRRC, the SNR is 18.6 dB.

Table 14.1 Matched-Filter Tap Numbers and Delay vs. Output SNR

No. Taps of the Matched Filter	481	241	121	61	31	21
Delay introduced (micro seconds)	5.575	2.788	1.394	0.697	0.348	0.232
Matched-filter output SNR (dB)	58.1	50.8	42.3	25.7	18.6	17.1

14.2.1.4 Synchronizing Transmitted Signal with Received Signal

It is necessary to frequency-synchronize the transmitted signal with received signal. If there is any frequency difference between the signal from the main transmitter and the EDOCR output signal, it will result in an apparent Doppler effect, which can be fatal to the performance of legacy DTV receivers.

There are two methods to achieve synchronization. One is to use an external frequency reference signal such as GPS; the other is to use the received signal from the main transmitter. The EDOCR does not use an external reference signal, but uses frequency offset information with respect to the received signal. When performing carrier and timing recovery at the demodulator, frequency offset information can be obtained. This information is used directly at the re-modulator for carrier synchronization.

By using this method, the EDOCR output signal can maintain frequency synchronization with the signal received from the main transmitter. Although the phase of an EDOCR output may not be equal to the phase of the received signal from the main transmitter, that fact will not have much impact on signal reception. The phase noise of the (pilot) carrier output from an EDOCR should be less than -104dBc/Hz at 20 kHz offset from the carrier frequency. This condition can be satisfied by adjusting the loop filter bandwidth of the carrier and timing recovery block.

14.2.2 Signal Processing Subsystem

The signal processing subsystem includes the following elements: 1) equalizer and intelligent slicer, and 2) RF Watermark insertion system.

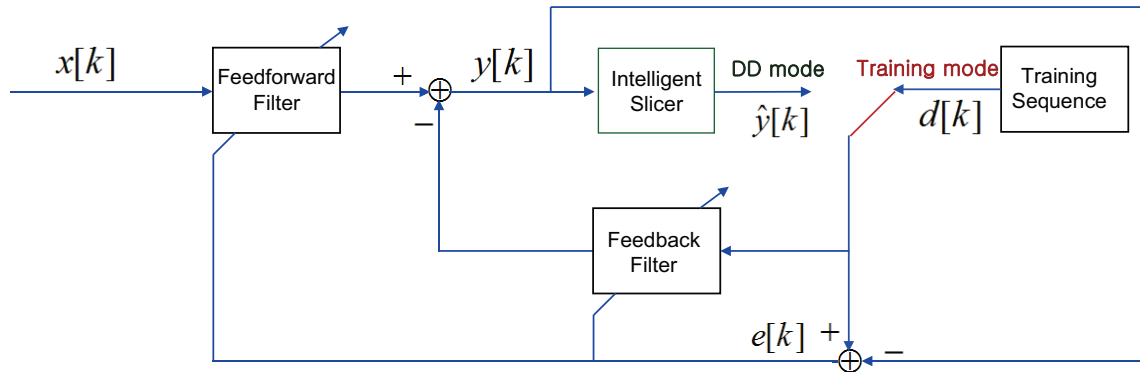


Figure 14.3 Equalizer with an intelligent slicer for the Baseband Equalization DOCR.

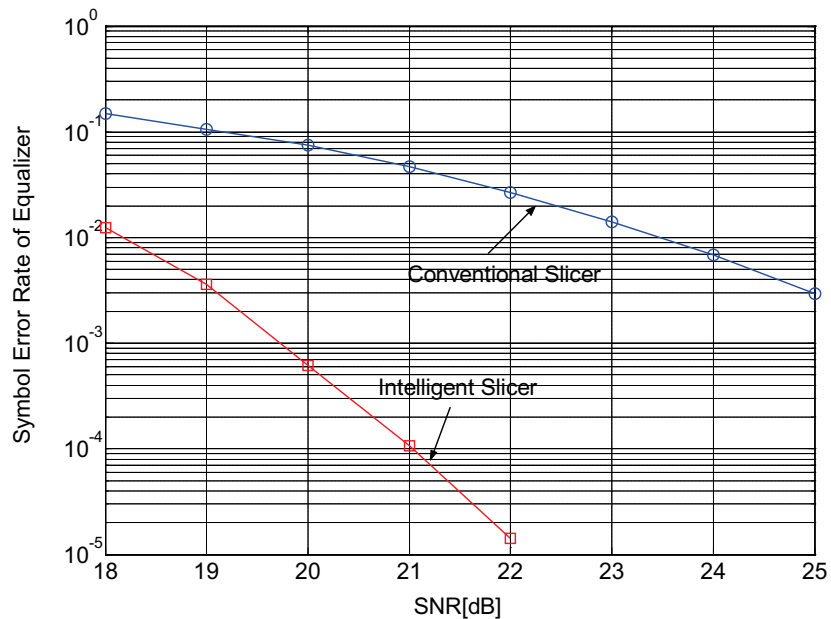


Figure 14.4 Comparison of the conventional slicer and the intelligent slicer.

14.2.2.1 Equalizer with Intelligent Slicer

The equalizer in the EDOCR consists of a 40-tap feedforward (FIR) filter, a 128-tap feedback (IIR) filter, and an intelligent slicer, as shown in Figure 14.3. Equalization is used to compensate for various forms of linear distortions. It suppresses inter-symbol interference in the received signal caused by multipath distortion, output signal loopback, adjacent channel interference, and imperfect matched-filtering. The intelligent slicer, which has the capability of error correction, improves SNR vs. BER performance.

Since DOCR internal delay can seriously affect the performance of ATSC legacy receivers, it is desirable that equalizer processing delay be minimized. This means that the equalizer should have the minimum delay possible in the feedforward filter. The delay is determined by the position of the reference tap, which allocates the anti-causal part and the causal part of the feedforward filter. In the EDOCR design, considering the tradeoff between equalization

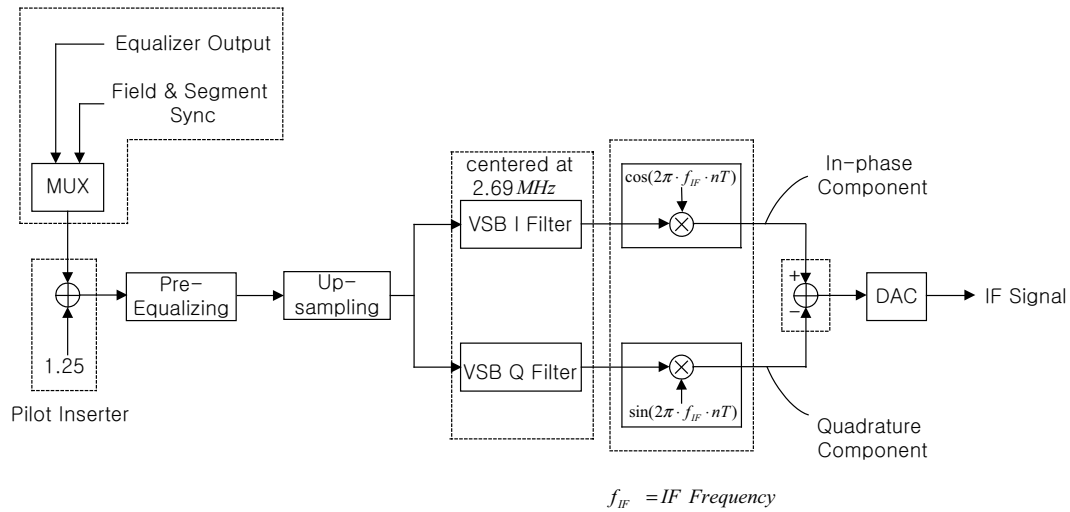


Figure 14.5 VSB re-modulator for the Baseband Equalization DOCR.

performance and delay, the reference tap of the feedforward filter is placed at the 6th position. The equalization delay is thus 5 symbol times.

The use of the intelligent slicer further improves EDOCR performance. Figure 14.4 shows the performance improvement of the intelligent slicer over a conventional memoryless slicer under Brazil A channel conditions. It is important to note that the intelligent slicer improves the correct-decision probability, or SER performance, by more than 5 dB.

Since the equalization in an EDOCR also can compensate for antenna loopback effects, high antenna isolation, which has been the most difficult problem in implementing DOCR systems, can be obtained.

14.2.2.2 RF Watermark Insertion

Since the equalization process regenerates a clean baseband signal, the RF Watermark, which behaves as low level noise on the signal, is simultaneously eliminated. An EDOCR system has the capability of either re-inserting the original RF Watermark from the received signal from the main transmitter or inserting a new RF Watermark to identify itself. In the first case, the EDOCR operates like an IF or RF processing DOCR, where the main transmitter RF Watermark is simply passed through. In the second case, the EDOCR operates similarly to a distributed transmitter.

14.2.3 Transmitting Subsystem

EDOCR system delay is an important parameter in an SFN network design. Reduction of the time delay is also crucial in the transmitting subsystem in the EDOCR. The time delay of a VSB filter depends upon the number of its taps, which is limited by the required RF spectrum mask.

14.2.3.1 Re-modulator and Pre-Equalizer

VSB re-modulation is performed with a VSB filter pair consisting of a 141-tap in-phase and a 141-tap quadrature windowing VSB filter (see Figure 14.5).

In a conventional transmitter, the VSB filter is made up of an SRRC filter, and the number of taps is determined to satisfy the transmission spectrum mask. If the up-sampling rate for VSB

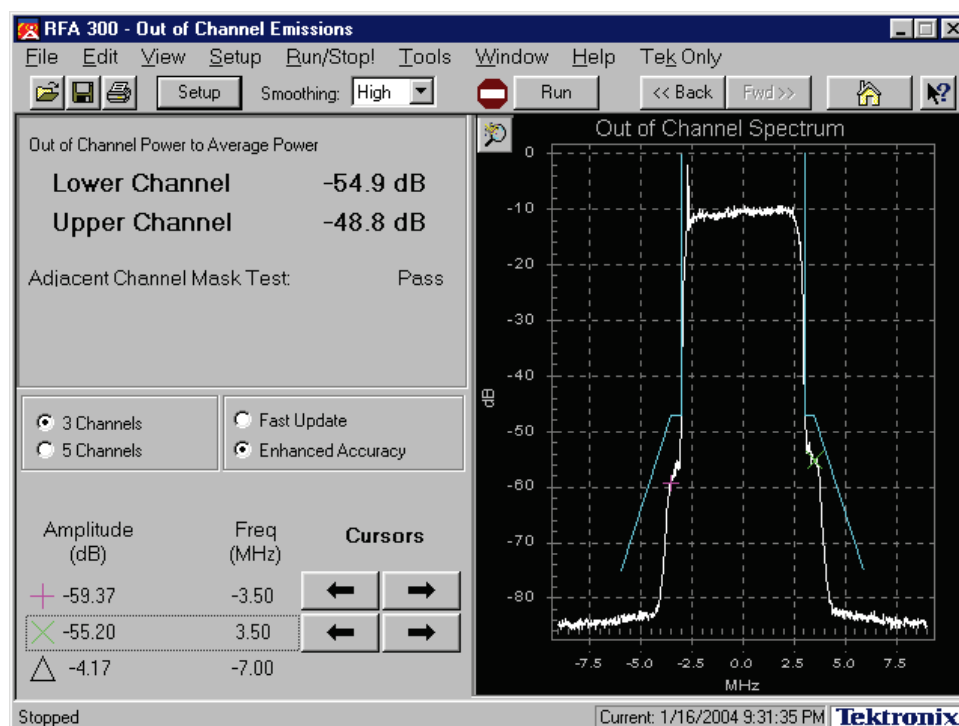


Figure 14.6a Transmission spectrum of a VSB signal after filtering by an SRRC filter with 521 taps and an RF channel-filter.

filtering is 4, the SRRC filter requires more than 500 taps, which means almost 6 μs delay. In an EDOCR re-modulator, however, since time delay is critical, an SRRC filter with such a large number of taps is prohibited. It has been demonstrated that, for VSB modulation, the windowing VSB filter with about 140 taps is adequate to meet RF emission mask requirements. Figures 14.6a and b show the spectra of the transmitter output signals generated by a 521-tap SRRC filter (6.04 μs delay) and a 141-tap windowing VSB filter (1.626 μs delay) after channel filtering. Both signals satisfy the spectrum mask.

The cost for reducing the delay by using a windowing VSB filter is a moderate reduction of the SNR of the re-transmitted signal because the windowing VSB filter is not an ideal Nyquist pulse-shaping filter¹⁰. Simulations show that an SRRC filter with 521 taps achieves an SNR = 42.3 dB, while a windowing VSB filter with 141 taps provides an SNR = 30.6 dB. (Note: in the simulation, the channel was assumed to be ideal, and the number of taps of the SRRC matched filter in the receiver was set to be 121.) Although there is about a 12 dB SNR drop, the decrease in SNR can be partially compensated with the pre-equalizer, which is generally used for the compensation of in-band distortion caused by the emission masking filter. A pre-equalizer can also play the role of restoring non-ideal pulse-shaping. Table 14.2 shows the pre-equalization performance with a 101-tap finite impulse-response filter with reference-tap positions of 10, 15, and 20, in conjunction with a 141-tap (4 times over-sampling) windowing VSB filter. Table 14.3 lists total system delay. Since the SNRs of the filtered signals are greater than the required 27 dB,

10. A Nyquist pulse-shaping filter is a pulse-shaping filter that satisfies the condition of causing no ISI.

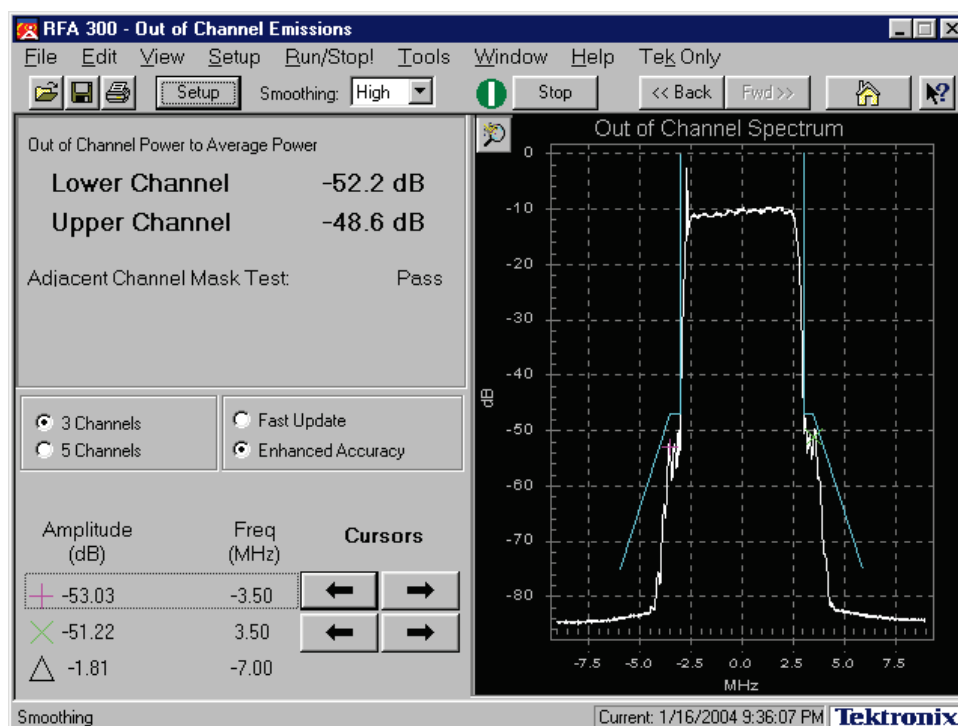


Figure 14.6b Transmission spectrum of a VSB signal after filtering by a windowing VSB filter with 141 taps and an RF channel-filter.

they all meet the ATSC system emission requirement. A pre-equalizer with a reference-tap position of 10 is sufficient for re-transmission.

Table 14.2 SNR Values vs. Position of the Reference Tap of the Pre-Equalizer

Position of Reference Tap of the Pre-Equalizer	10	15	20
Output SNR [dB]	33.7	33.8	41.5
Pre-equalization delay (microseconds)	0.836	1.301	1.765
141 windowing VSB filter delay (microseconds)	1.626	1.626	1.626
Total re-modulation delay (microseconds)	2.462	2.927	3.392

Table 14.3 Total System Delay of EDOCR

Subsystem	Delay Time (microseconds)
Demodulator	0.697
Equalizer	0.465
Modulator	2.462
A/D	0.200
D/A	0.500
Analog Part	0.450
Implementation	0.200
Total Delay	4.974

14.2.3.2 Up Converter

The IF VSB-modulated signal is up-converted to the RF band that has exactly the same frequency as the input RF signal of the EDOCR. To maintain the carrier frequency synchronization with the main transmitter carrier frequency, the local oscillator used in the RF up-conversion is synthesized from the common reference source that was used in the down-conversion. Because there are harmonic components at the output of a digital-to-analog converter, they must be removed at the IF stage. This can be done by using a low-pass filter or a band-pass filter. Meanwhile, because the up-converted signal contains unwanted sideband spill-over, a band-pass filter is used for removing the unwanted signal.

14.2.3.3 High-power amplifier

A high power amplifier (HPA) is used for amplifying the up-converted RF signal. The gain of the HPA depends on the operating power of the EDOCR and its application; i.e., whether it is used as a gap filler or as a coverage extender. In the gap filling application, an EDOCR can be operated at under 10W. However in the coverage-extending application, it can be operated at over 1000W.

14.2.3.4 Channel filter

A channel filter is used for removal of adjacent channel spill-over in the transmitted signal. The channel filter used in an EDOCR must provide proper filtering to meet the RF emission mask specified by the spectrum authority.

14.3 Implementation of Baseband Equalization Distributed Translator Systems

This section provides a hardware implementation example of a Baseband Equalization Distributed Translator (EDTxR). The EDTxR system basically uses the same design principles as those of EDOCR system described in the previous section and shown in Figure 14.1. Unlike EDOCR, however, the input and output channels of EDTxR are different. Such a difference brings more flexibility in its design procedures, resulting in some improvements in its hardware implementation as compared to that of EDOCR.

In the following sections, the three subsystems (receiving, signal processing, and transmitting subsystems) of EDTxR are briefly explained and their differences with those of EDOCR are highlighted.

14.3.1 EDTxR Receiving Subsystem

In the EDTxR receiving subsystem, the structure and operation of the pre-selector, low-noise amplifier (LNA), and down-converter are the same as those for EDOCR explained in Sections 14.2.1.1 to 14.2.1.3. These modules are used to bring the input RF signal down to IF.

Concerning the demodulator for EDTxR, there are some differences with that of EDOCR (shown in Figure 14.2) in the way that it can use a filter with higher number of taps to achieve a better response with respect to Inter-Symbol Interference (ISI). The differences between the demodulator for the two systems are explained in more details in the following section.

14.3.1.1 EDTxR Demodulator

Like EDOCR, a square-root-raised-cosine (SRRC) filter can be used in EDTxR demodulator for matched-filtering. Increasing the number of taps of such filter increases the S/N performance and reduces ISI, but increases the delay and complexity of the filter as well.

In the case of EDOCR, to avoid a long internal delay, the number of taps was recommended to be between 31 and 61 for when the symbols are over-sampled at 4 times the ATSC system symbol rate (see Section 14.2.1.3). For EDTxR, since the internal time delay is not a critical problem, the SRRC filter can use a larger number of taps. Under such circumstances, however, it is the complexity of the filter that puts a restricting limit on the number of taps.

For EDTxR, the number of taps is recommended to be between 61 and 121. As shown in Table 14.1, the result of the S/N simulation under such conditions is between 25.7 and 42.3 dB.

It should be noted, however, that like EDOCR, the use of the baseband equalizer with intelligent slicer (see Section 14.2.2.1) can enhance the received S/N and can compensate for the distortions caused by imperfect matched filtering. Therefore, such number of taps along with the enhancing effect of the baseband equalizer with intelligent slicer can provide an almost ISI free conditions and a fairly high S/N at the output of the demodulator.

14.3.1.2 Frequency Synchronization and Emission Time Adjustment of Multiple EDTxRs

If a number of EDTxRs are forming a single frequency network (SFN), their output frequencies should be precisely synchronized. There should also be the possibility of controlling and adjusting the emission timing of such translators with respect to each other.

As in the case of EDOCR, two methods are proposed for achieving frequency synchronization among different EDTxRs receiving their input signals from a single source. One method would be to use GPS, and the other to use the input frequency of the translators (main transmitter signal) as the synchronization reference. It should be noted, however, that the difference between input and output frequency of a translator may make the hardware implementation of the frequency control segment of EDTxR different from that of EDOCR.

As far as the emission time adjustment of multiple EDTxRs is concerned, each translator should be capable of providing an adjustable additional delay to the signal to enable delay adjustments in the overlapping areas of the distributed translators.

14.3.2 EDTxR Signal Processing Subsystem

Signal processing subsystem of EDTxR, consisting of equalizer with intelligent slicer and RF watermark system, is the same as that of EDOCR explained in Section 14.2.2 and shown in Figure 14.3. However, since the internal time delay of EDTxR is not a critical issue, some modifications are possible for improving the performance of the equalizer. Such modifications consist of increasing the number of taps of the feedback IIR filter, and moving forward the position of the reference tap of the feedforward (FIR) filter used in the equalizer module.

For EDTxR, the number of taps of the IIR filter in the equalizer is recommended to be increased from 128 (for EDOCR) to 196. The position of the reference tap of the FIR filter, which was set to the 6th position for EDOCR to get a delay of no more than 5 symbol time, can also be moved forward to higher positions.

14.3.3 EDTxR Transmitting Subsystem

Except the re-modulator and pre-equalizer, all other elements of the transmitting subsystem of EDTxR (up-converter, high-power amplifier, and channel filter) are the same as those of EDOCR explained in Sections 14.2.3.2 to 14.2.3.4. It should be noted, however, that the output power limitation of EDOCR, which exists because of the presence of loop-back signal from (the same channel) output to input, does not apply for EDTxR.

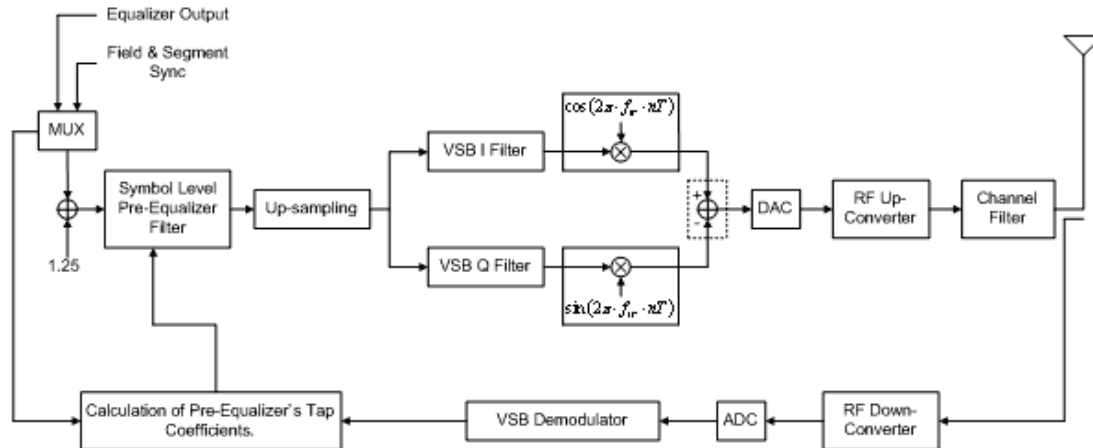


Figure 14.7 Re-modulator and pre-equalizer of EDTxR.

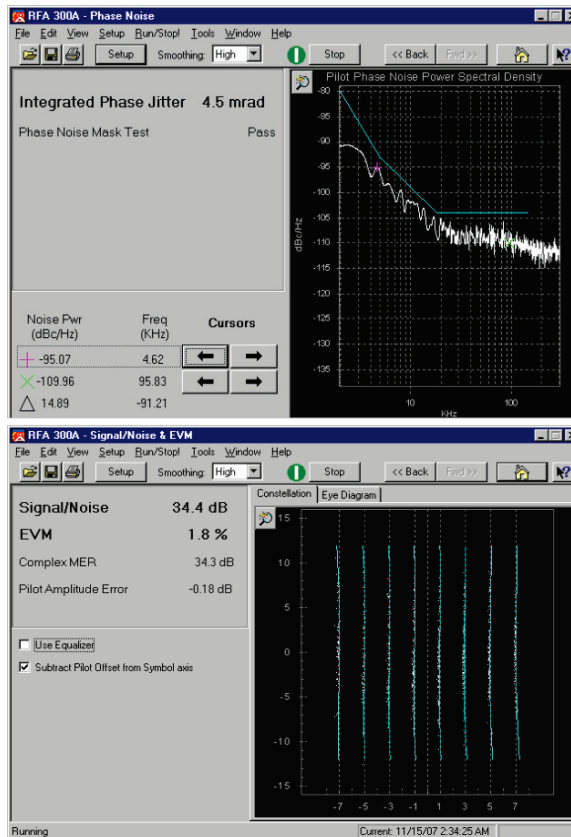
14.3.3.1 Re-Modulator and Pre-Equalizer for EDTxR

Figure 14.7 shows the re-modulator and pre-equalizer of EDTxR. By using the same design principles as those of EDOCR, re-modulation of the VSB signal in EDTxR is done in three steps. The first step is to up-sample the VSB signal to which TxID, field and segment synch, and pilot have already been added. Then the signal is filtered by a VSB I/Q pulse shaping filter. At this stage, the center frequency of both I/Q components will be 2.69 MHz. Finally, the filtered I/Q components are up-converted to the center frequency of the IF band (f_{IF}), and combined together to form the IF signal.

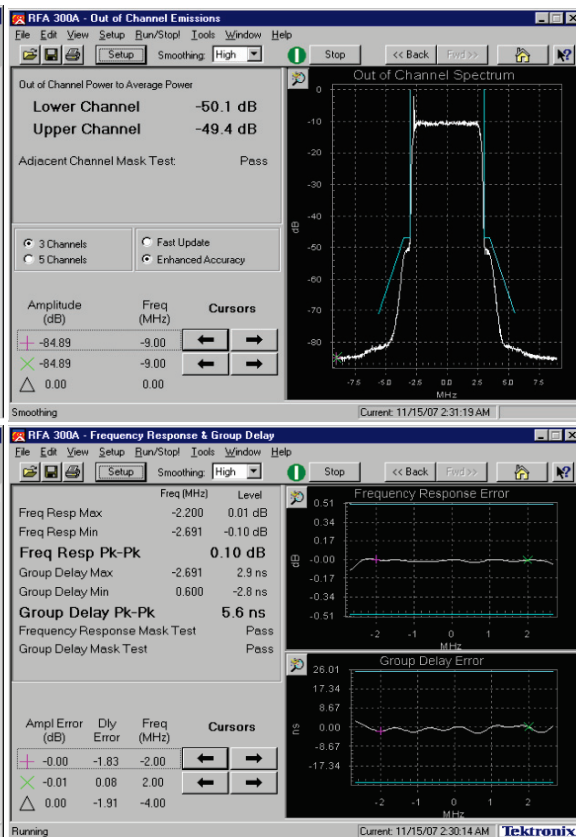
One of the differences between the re-modulator for EDTxR and the one for EDOCR is the type of the VSB I/Q pulse shaping filter used in this module. In the case of EDOCR, such filter was selected to be an equi-ripple (ER) filter, which has the capability of providing a good out-of-band suppression at low delays, but the disadvantage of allowing a lot of in-band ripples that results in reducing the output S/N. In the case of EDTxR, since the internal time delay is not a critical factor, square root raised cosine (SRRC) filter can be used instead of ER filter. Such filter (SRRC) is normally used as the VSB filter in a conventional transmitters and the number of its taps is determined to satisfy the transmission spectrum mask. The VSB re-modulated signal must maintain a S/N greater than 27 dB, and a shoulder amplitude of more than 47 dB with respect to total average DTV power (36 dB with respect to the DTV spectrum flat top) to meet the emission mask requirements. It can be shown that the VSB pulse shaping SRRC filter should theoretically have more than 420 taps to simultaneously meet the output S/N of above 27 dB and shoulder amplitude of greater than 47 dB.

It should be noted, however, that the output of EDTxR after the high-power amplifier should also comply with the emission mask requirements. To meet such requirement, a channel filter is required after the power amplifier (see Figure 14.7). But a channel filter with a good out-of-band suppression capability causes a lot of in-band group delay which degrades the output signal quality by reducing its S/N. To compensate for such degradation, a symbol level pre-equalizer filter is used before up-sampling the VSB signal (see Figure 14.7). The coefficients of such filter are continuously updated by comparing a demodulated sample signal from the output of the translator, and the input signal to the re-modulator (the baseband signal).

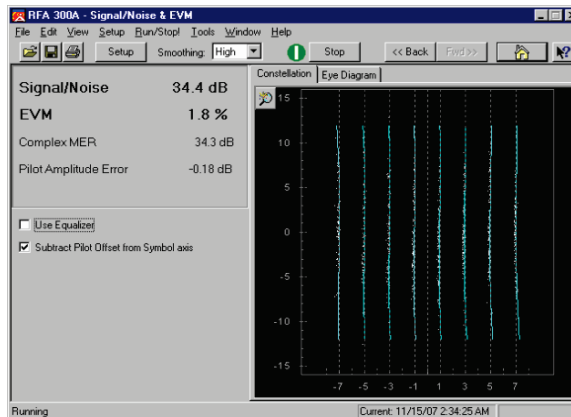
(a) Phase Noise



(b) Spectrum



(c) Signal Constellation



(d) Group Delay and Frequency Response

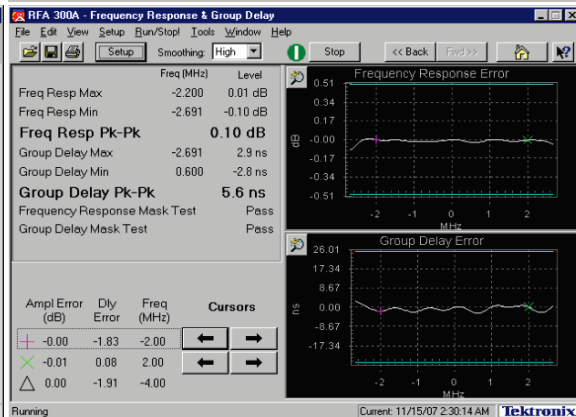


Figure 14.8 Phase noise, spectrum, signal constellation, and group delay and frequency response of an EDTxR output signal.

For EDOCR, a 101-tap finite impulse response filter with reference tap position of 10 was recommended to be used for the pre-equalizer. For EDTxR, since the quality of the signal is already higher due to the use of SRRC instead of equi-ripple filter, the number of the taps of the pre-equalizer can be reduced. For the example EDTxR hardware implementation, the number of taps of such filter was selected to be 64.

Figure 14.8 shows the phase noise, spectrum, signal constellation, and frequency response and group delay of an EDTxR output signal that uses a 481-tap SRRC filter as its VSB pulse shaping filter and a 64-tap FIR filter as its pre-equalizer.

14.4 Uses of DTx Technologies for Other Purposes

Distributed Transmission technology incorporates certain functionality that can be applied beneficially in other applications involving multiple or even single transmitters that have been designed with the DTx technology. This functionality includes transmitter identification and transmitter remote control. Other uses of the technology are radiolocation and synchronization of Enhanced VSB (E-VSB) Signal Processing.

14.4.1 DTV Transmitter Identification by the RF Watermark

There is a need to identify the individual transmitters in a Distributed Transmission Network that are actually received anywhere in the network in order to confirm network design and to optimize performance. Identification is accomplished using a low level RF Watermark that is transmitted with the main signal. The RF Watermark is a spread spectrum signal “buried” under the main signal. Without the RF Watermark, there would be no way to identify individual transmitters other than by turning them on and off. Determining the identification of a received signal by demodulation is not possible when all transmitters carry the same synchronous bit stream. The RF Watermark can be detected and its code identified when the RF Watermark signal is over 50 dB below the signal from another transmitter. Test equipment can be designed to provide for rapid demodulation and detection of the RF Watermark codes in the field.

It is equally important for all DTV stations, even those with only one transmitter, to have the ability to identify interfering signals and their received signal levels. This will become crucial as all DTV stations increase power to maximized levels and the spectrum is repacked. When DTV stations interfere with one another, there is no easy way to know whether the reason receivers in an area have stopped working is due to a fault in the signal from the desired station or due to interference. Moreover, DTV interference to NTSC signals appears as unidentifiable white noise. In both cases, without using the RF Watermark, identifying interfering DTV stations could only be practically accomplished by turning transmitters on and off.

14.4.2 DTV Transmitter Remote Control and Monitoring

The RF Watermark signal may be modulated to carry slow speed (≈ 40 bps) data communications from one or more transmission sites to convenient monitoring locations. Data communication from the studio to the transmitters can be via the STL to a single transmitter or to a Distributed Transmission Network. The RF Watermark provides a return link for telemetry that complements the control data that can be sent over the STL. The Remote control of transmitters can include traditional functions and can be used for a single traditional DTV transmitter until additional transmitters are added.

14.4.3 DTV Signals with RF Watermark for Radiolocation

The RF Watermark signals can be applied for location finding (radiolocation), where the signals of three or more DTV transmitters carrying RF Watermark signals can be used for calculating the accurate geographic location of the RF Watermark receiver. In comparison with other candidate terrestrial communication systems used for radiolocation, such as third generation cellular phone systems and other terrestrial microwave systems, for location-finding applications, the DTV signal with RF Watermark has a number of relative benefits. These benefits include wider bandwidth, larger coverage area, much longer RF Watermark sequence, and, most importantly, lower operating frequency for better building penetration capability. All of these characteristics will make the DTV signal with RF Watermark a better location finding system, especially for indoor location finding, where the GPS system does not work.

14.4.4 Synchronizing E-VSB Signal Processing

Implementation of E-VSB requires synchronization and processing of multiple input streams to a transmitter or transmitter system. These multiple input streams carry the E-VSB and legacy 8-VSB payloads. In a DTx system, these functions must be performed in the DTxA so that all transmitters in a network will synchronize to its output. The DTxA will multiplex its input streams

and produce an MPEG-2 Transport Stream output stream carrying the multiplexed 8-VSB and E-VSB packets, and a field rate side channel that contains the field sync reserved bits used by the E-VSB system.

A compliant DTxT inherently will be capable of E-VSB operation; the additional burden of E-VSB operation falls entirely upon the DTxA. For this reason, it may be desirable for both transmitter manufacturers and broadcasters to use a DTxA and a DTxT when operating a single transmitter system.

The advantage to the broadcaster is that a single MPEG-2 Transport Stream stream may be formed that contains all of the E-VSB data. That single stream may be transmitted over a conventional STL, as opposed to carrying separate E-VSB and 8-VSB streams to a transmitter. In other words, the E-VSB multiplexing operation is carried out at the studio rather than at the transmitter site.

A/111, Annex A:

Distributed Transmission Design Examples

A.1 INTRODUCTION

In designing Distributed Transmission Networks, whether they contain Distributed Transmitters (DTxTs), Distributed Translators (DTxRs), or Digital On-Channel Repeaters (DOCRs), there are a number of evaluation steps that must be undertaken if the performance expected of the network is to be understood and optimized. The various steps themselves were described in Section 11 of this Recommended Practice. This annex provides several concrete examples of the application of the techniques described above, with emphasis on addressing some of the different sorts of situations that arise in real world designs. Indeed, the examples described below are real world designs undertaken for actual DTV stations in the locales described. Two of the cases are not finalized or fully optimized designs but nevertheless serve as good examples of the sorts of analyses required for such cases. In one case, the system described has already been partially implemented and daily is delivering DTV signals to consumers who would not be receiving any over-the-air DTV without its use. It should be noted that the examples given deal only with the issues of network internal interference and do not treat considerations of interference to and from other stations, which would be required in a complete network design process.

A.2 EXAMPLE CASES

The examples, in order, are of a fully maximized DTV facility that nevertheless is cut off by terrain from all the population centers within its service contour, a major city in flat earth region where multiple transmitters may be used in place of a single central transmitter facility, and a small city shielded by a mountain where increased signal levels will allow set top reception that otherwise would not be available. The objective of this annex is not to provide detailed design criteria and processing results but rather to examine presentations of the kinds of results that can be expected during the design and optimization processes.

A2.1 Example 1 — Terrain-Shielded Population Centers

Example 1 is a station in Central Pennsylvania (WPSX-DT) on DTV Channel 15. Its NTSC transmitter is located in Clearfield, PA, and its DTV reference facilities were in the same place. There are three major population centers in the station's service area: State College, Altoona, and Johnstown, all in Pennsylvania. Even with a fully maximized facility at Clearfield, it is not possible to provide more than spotty, outdoor antenna service in the three population centers because of terrain blockage by a ridge called Rattlesnake Mountain. Attempts to increase the antenna height at Clearfield were not successful because of a tight mesh of commercial and private aircraft routes in the region. Moving the transmitter near one of the population centers was not really an option because the population centers are obstructed from one another so that all three could not have been served from a location near any one of them, and such a move would have eliminated service in the northern tier of the station's service area — a rural area where it is often the only over-the-air television service. Consequently, a decision was made to install a fully maximized facility at the Clearfield site to serve the rural communities to the north, including

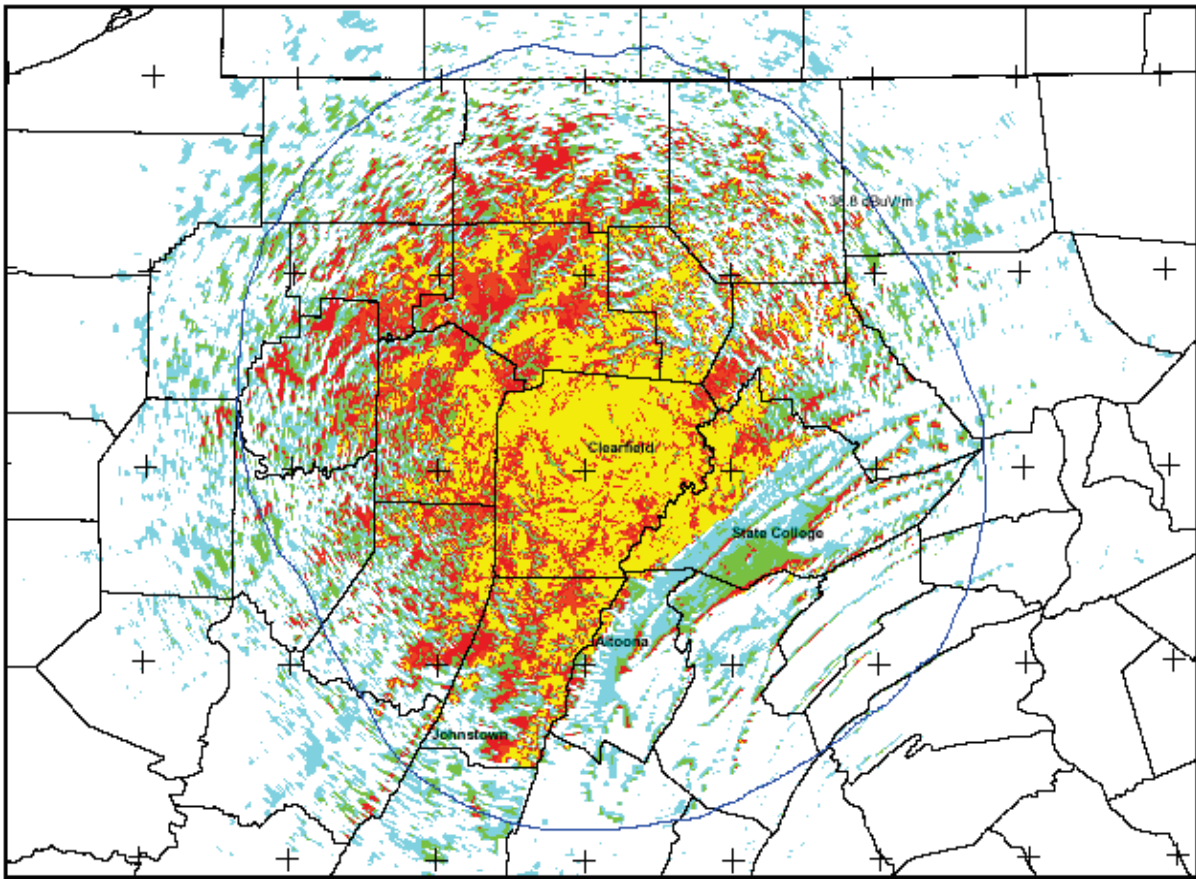


Figure A.1 WPSX-DT with maximized facilities at Clearfield, PA.

those north of Rattlesnake Mountain but south of the transmitter, and to supplement it with a series of smaller Distributed Transmitters (DTxTs) to serve each of the population centers.

The initial problem and some of the results of the solution are shown in the maps and tables that follow. Figure A.1 shows the field strengths predicted from the fully maximized facility at Clearfield with the locations of the three population centers indicated. The color coding of the Longley-Rice propagation model results in Figure A.1 and in all comparable maps throughout this annex are >80 dBu in yellow, 70–80 dBu in orange, 60–70 dBu in red, 48–60 dBu in green, and 41–48 dBu in cyan. Generally, the yellow areas can be considered to have highly reliable indoor antenna reception, the orange and red areas to have indoor reception with very good and good indoor antenna reception reliability, the green areas to have reliable outdoor antenna reception, and the cyan areas to have spotty (50 percent of locations) outdoor antenna reception. It should be apparent that the population centers have no possibility of indoor reception predicted and, in fact, offer only spotty outdoor antenna reception expectations. The areas with high signal levels (yellow, orange, and red) offer the potential for indoor reception in regions of state forest lands and game preserves (thus providing superb digital television service to the bears, deer, foxes, squirrels, and other inhabitants of those areas).

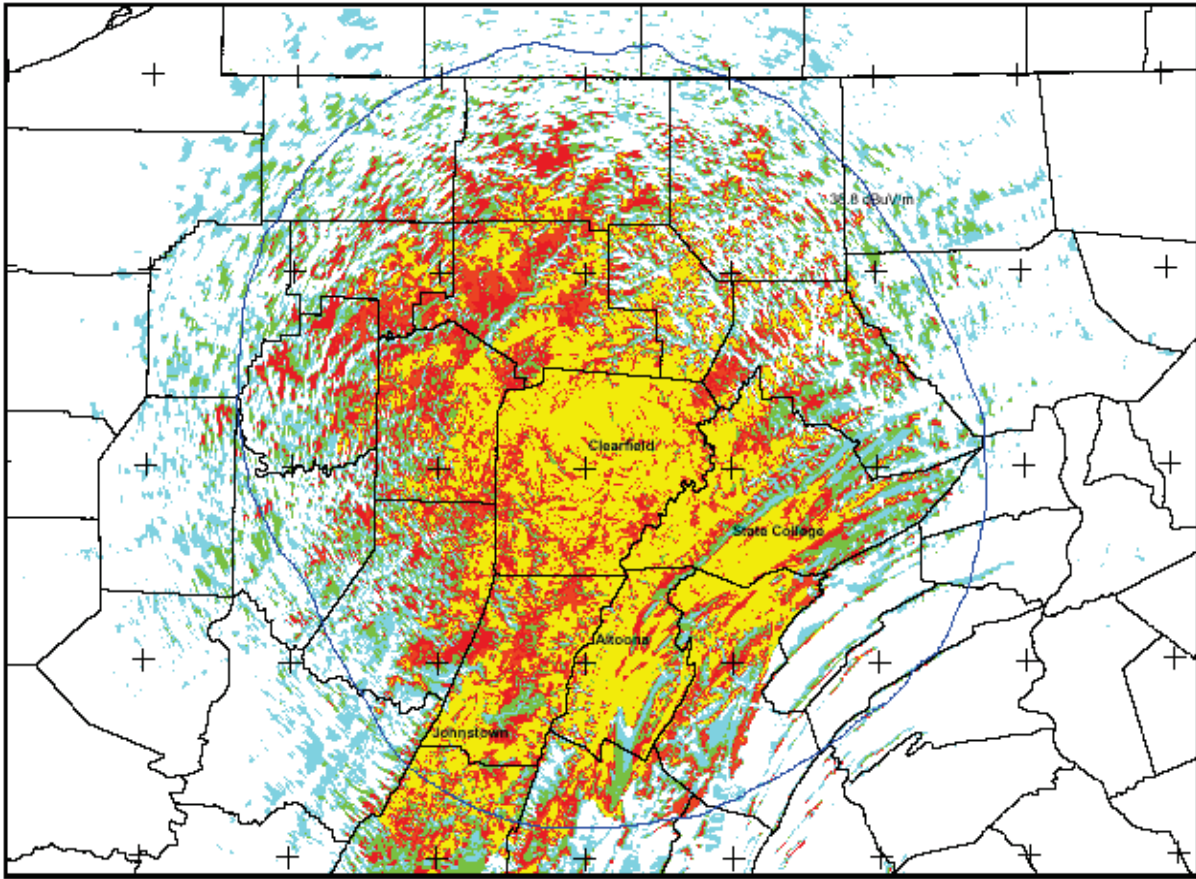


Figure A.2 WPSX-DT with three added Distributed Transmitters covering State College, Altoona, and Johnstown, PA.

The objective in the first example case was to provide adequate signal levels for indoor antenna reception in the population centers. This was accomplished through the addition of three DTxTs to the maximized facility in Clearfield. The Clearfield transmitter, of course, also operates as a DTxT in order to synchronize with the other transmitters. One added DTxT covers State College with a 50 kW ERP signal from a place called Pine Grove Mills. Two 25 kW ERP DTxTs cover Altoona and Johnstown. The Longley-Rice results with all four DTxTs is shown in the map of Figure A.2. The color coding is the same as in Figure A.1. It is evident from the map that virtually all of the three population centers receive signal levels adequate to enable highly reliable indoor antenna reception.

To examine the impact of the additions of the three lower power DTxTs, the population receiving each of the field strengths included in the color codes of Figures A.1 and A.2 is enumerated in Table A.1. In the table, the population numbers are cumulative from left-to-right; i.e., the population counted as receiving 70–80 dBu includes those receiving >80 dBu, and so on. It can be seen in the table that, when the lower power DTxTs are added, the population receiving one of the three higher field strength ranges, corresponding to those who are likely to have good

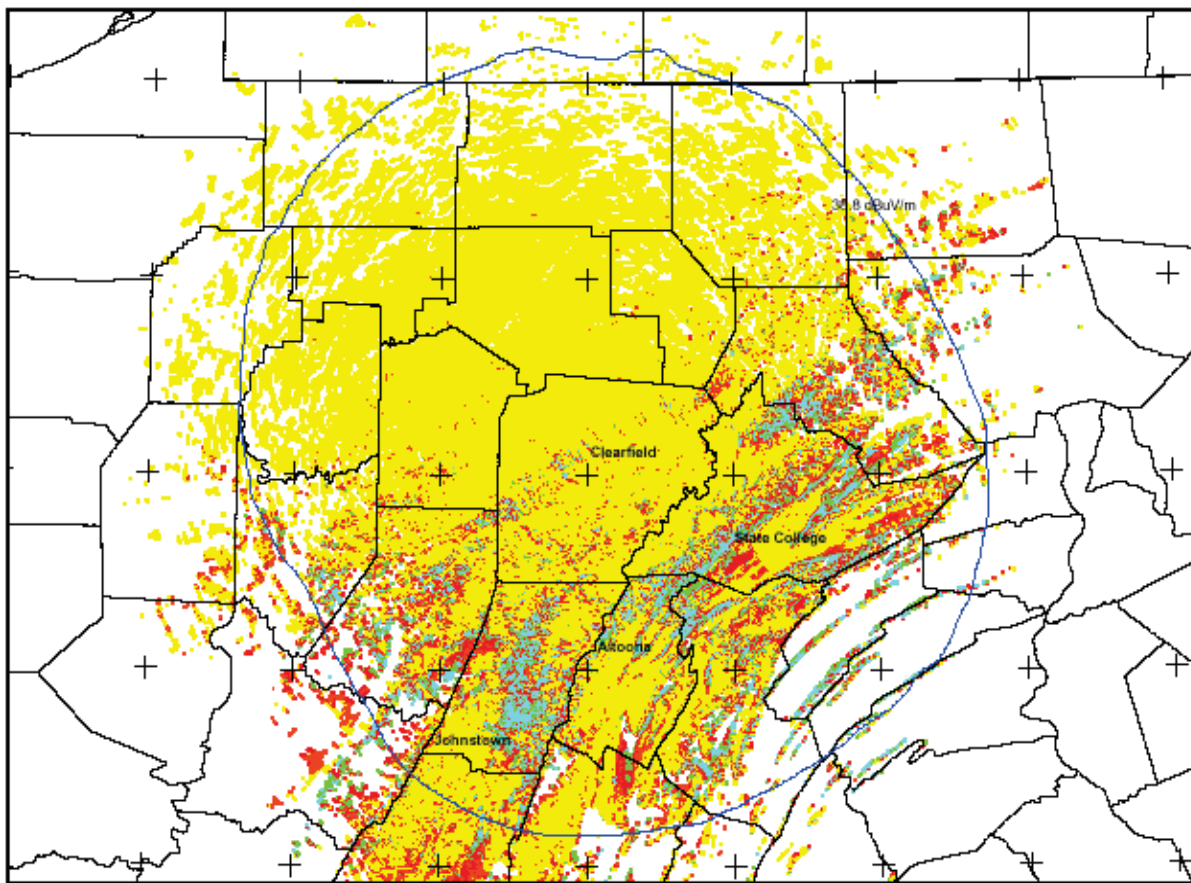


Figure A.3 $C/(I+N)$ between four Distributed Transmitters in WPSX-DT DTx network.

indoor antenna reception, multiplies by a factor of 2.5 to 4 compared to the population reached at those levels by the larger facility alone.

Table A.1 Population Reached by Transmitters in WPSX-DT Distributed Transmission Network

Transmitter	>80 dBu	>70 dBu	>60 dBu	>50 dBu	>39 dBu
Clearfield (810 kW)	109,075	158,833	242,365	416,410	797,388
State College (50 kW)	83,293	96,432	119,847	152,266	243,474
Altoona (25 kW)	111,278	134,750	165,535	259,954	441,834
Johnstown (25 kW)	87,216	107,980	135,005	184,972	264,632
Combined	384,853	471,945	598,655	750,777	1,044,701

The next level of examination of the system finds those locations in which the $C/(I+N)$ ratio falls below 20 dB. The areas identified are those in which consumer receiver adaptive equalizers will be active in dealing with apparent echoes coming from alternate transmitters. Figure A.3 is a map that identifies those areas with $C/(I+N) > 20$ dB in yellow, 15–20 dB in orange, 10–15 dB in red, 5–10 dB in green, and 0–5 dB in cyan. In this particular figure, an omnidirectional receiving antenna is assumed so that the worst case conditions at any location are identified. The further the

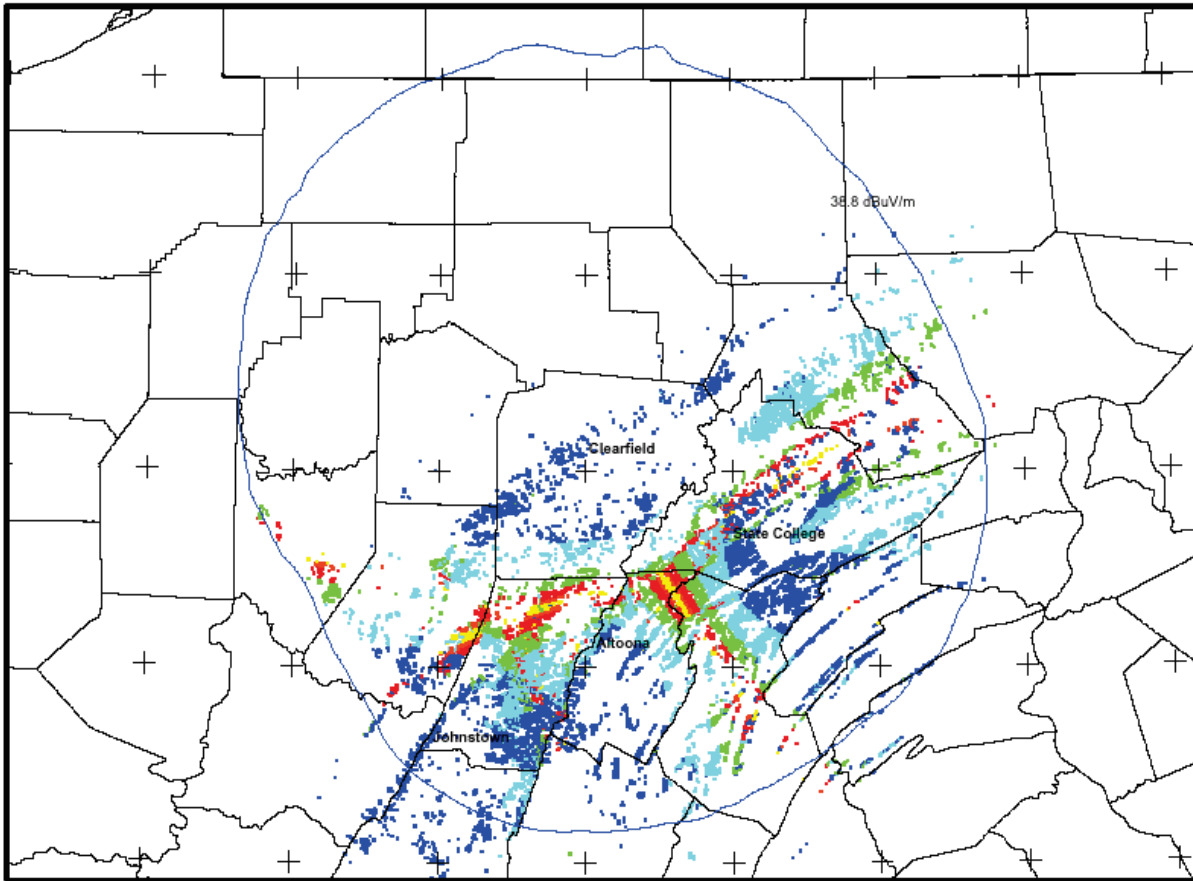


Figure A.4 Differential signal arrival times from transmitters in WPSX-DT Distributed Transmitter network.

combination of signals deviates from the yellow color coding, the harder receiver adaptive equalizers will have to work. The areas at the low end of the $C/(I+N)$ scale are areas where low delay spread timing will be important in order to keep adaptive equalizers operating in ranges where they can process the combination of signals likely to be received. It should be recognized that, because of the use of omnidirectional receiving antennas in preparing the map, it very often will be possible to improve on the values shown through the use of even moderately directional receiving antennas.

The differences in arrival times of the signals from the several transmitters are shown in Figure A.4. In this figure, the arrival time mapping is masked with a threshold of 20 dB $C/(I+N)$ so that only those areas in which consumer receiver adaptive equalizers are expected to be active are displayed. Again, omnidirectional receiving antennas are used to detect worst case conditions at each location with respect to $C/(I+N)$. The color coding scheme shows delay spreads from 0–5 μ s in yellow, 5–10 μ s in orange, 10–25 μ s in red, 25–50 μ s in green, 50–100 μ s in cyan, and >100 μ s in blue. The particular timing arrangement shown is not optimized from the standpoint of population receiving signals arriving with the various time differentials; instead it has been set to make the time differential scale more visible on the map.

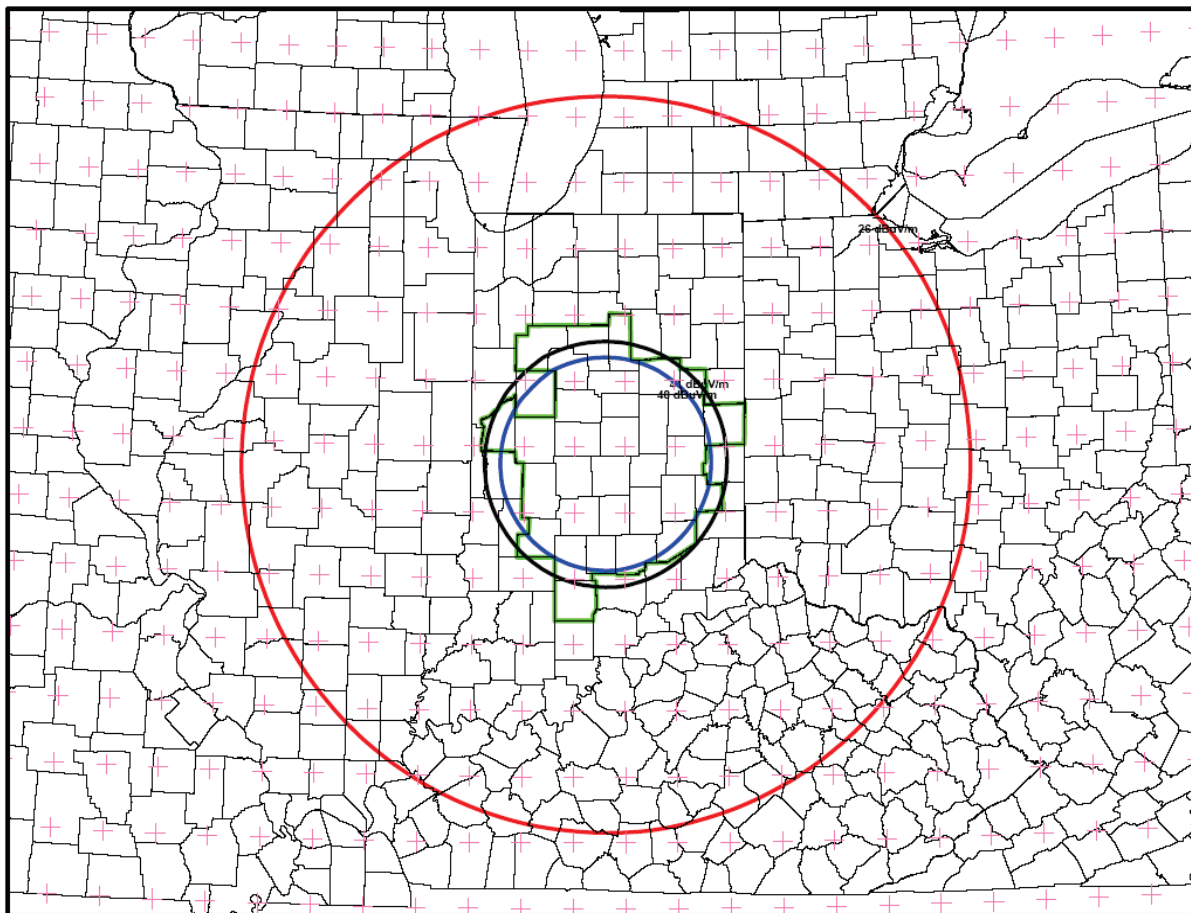


Figure A.5 Contours of maximized facility in Midwest with 1 MW at 365 meters.

A2.2 Example 2 — Flat-Earth Region and Interference Zone Comparisons

Example 2 is a city in the Midwest situated in relatively flat country where terrain shielding is not available. Because similar terrain extends for very long distances in nearly all directions, it should be one of the places where the method of propagation prediction using the FCC curves is most accurate. Therefore, it provides an opportunity for a case study of the relationship between the service contour and the interference contour of a station. The Midwest city also has two smaller cities nearby that have their own television stations and associated towers. The towers in the central and satellite cities provide good locations to serve as the starting points for a network design.

Starting with a look at the relationship between service and interference contours, Figure A.5 shows the Noise Limited (41 dBu – in black) and City Grade (48 dBu – in blue) contours (under the U.S. FCC rules) for a station operating in the UHF band with 1 MW ERP at an elevation of 365 meters height above average terrain (HAAT). The contours approximately circumscribe the Designated Market Area (DMA) in which the stations are located, indicated by the boundary line in green. These contours were developed using radials every one degree around the compass and the FCC F(50,90) curves used to predict DTV service areas. Also on the map of Figure A.5 is an interference contour, shown in red, calculated using the FCC F(50,10) curves at a field strength of

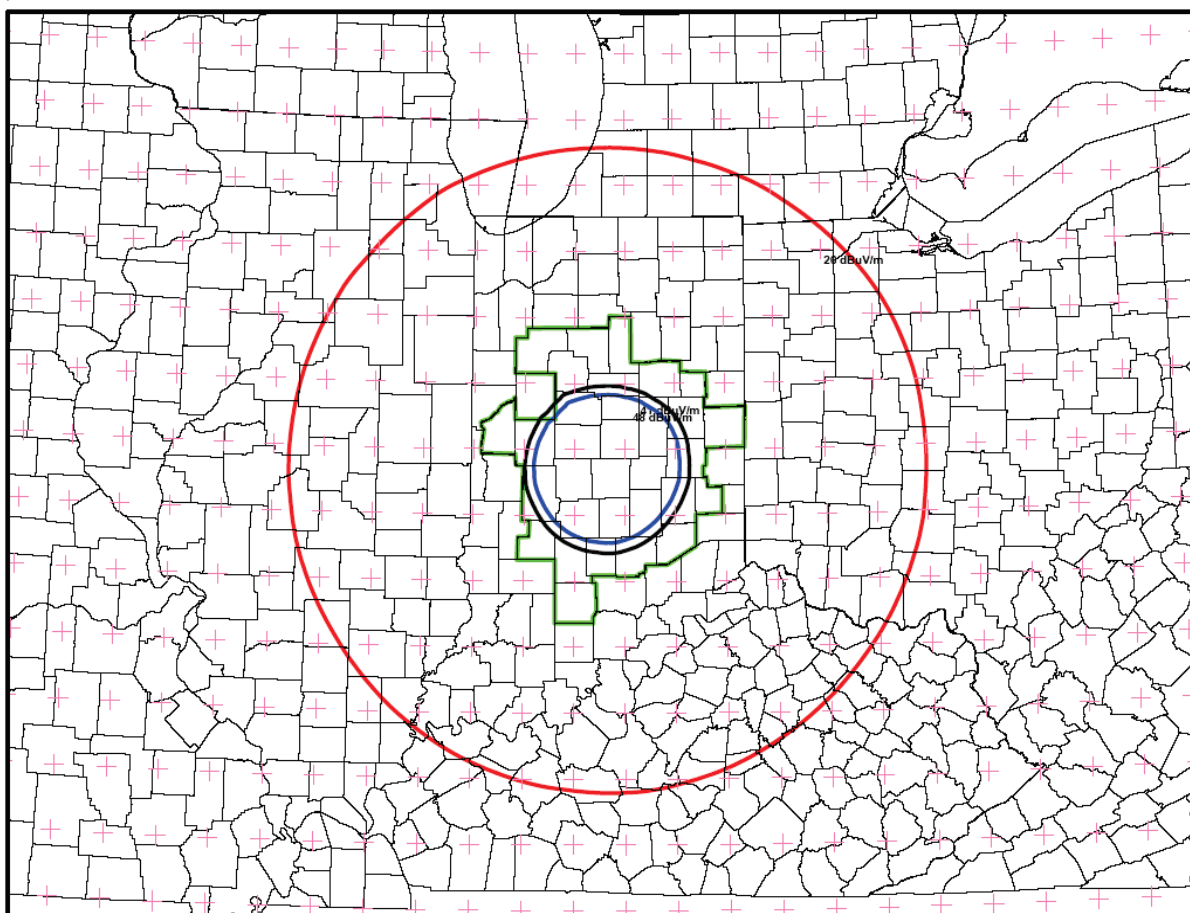


Figure A.6 Contours of Midwest facility with 1 MW at 85 meters above ground level.

26 dBu. 26 dBu is the threshold of interference to another DTV signal obtained when the co-channel D/U ratio of 15 dB is applied to the 41 dBu noise limited reception threshold planning factor. It should be evident from Figure A.5 that the interference contour extends about three times the distance that the noise limited service contour does.

Reducing the height of the antenna to about 85 meters above ground level while maintaining the 1 MW power level has the expected effect of reducing the service area and the corresponding interference area, as shown in Figure A.6, but notice the still very large interference contour. Even at the reduced antenna elevation, the interference contour still has about three times the radius of the service contour.

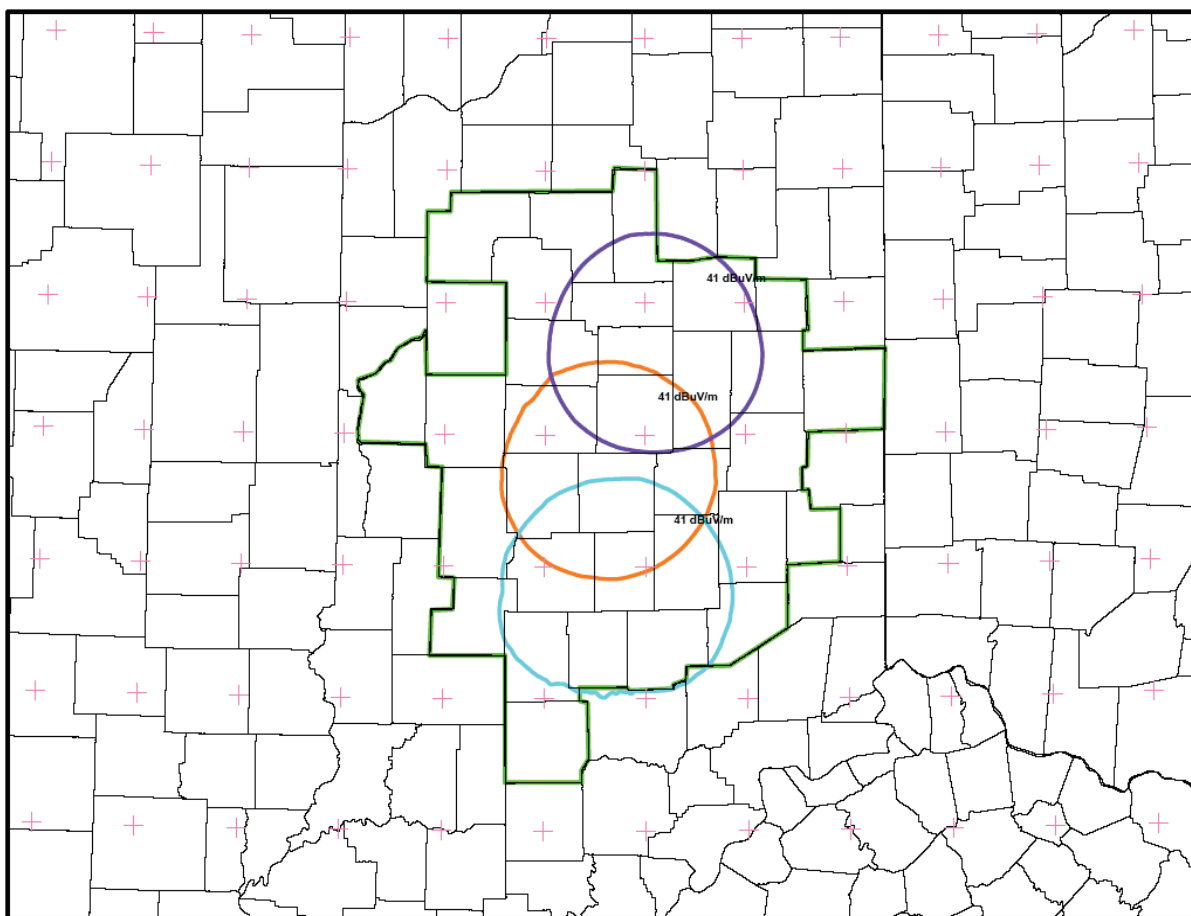


Figure A.7 Service contours of three transmitters each with 50 kW at about 75 meters above ground level.

Turning to what might be done with lower power Distributed Transmitters in this region, shown in Figure A.7 are the service contours of a group of three transmitters, each having 50 kW ERP and antenna elevations on the order of 80 meters above ground level, located on the towers at the three existing sites. By comparison with the service contour of Figure A.5, it can be seen that an area can be served by the three lower power transmitters approximating the area that could be served by a transmitter with much higher power from a single antenna at a much higher elevation.

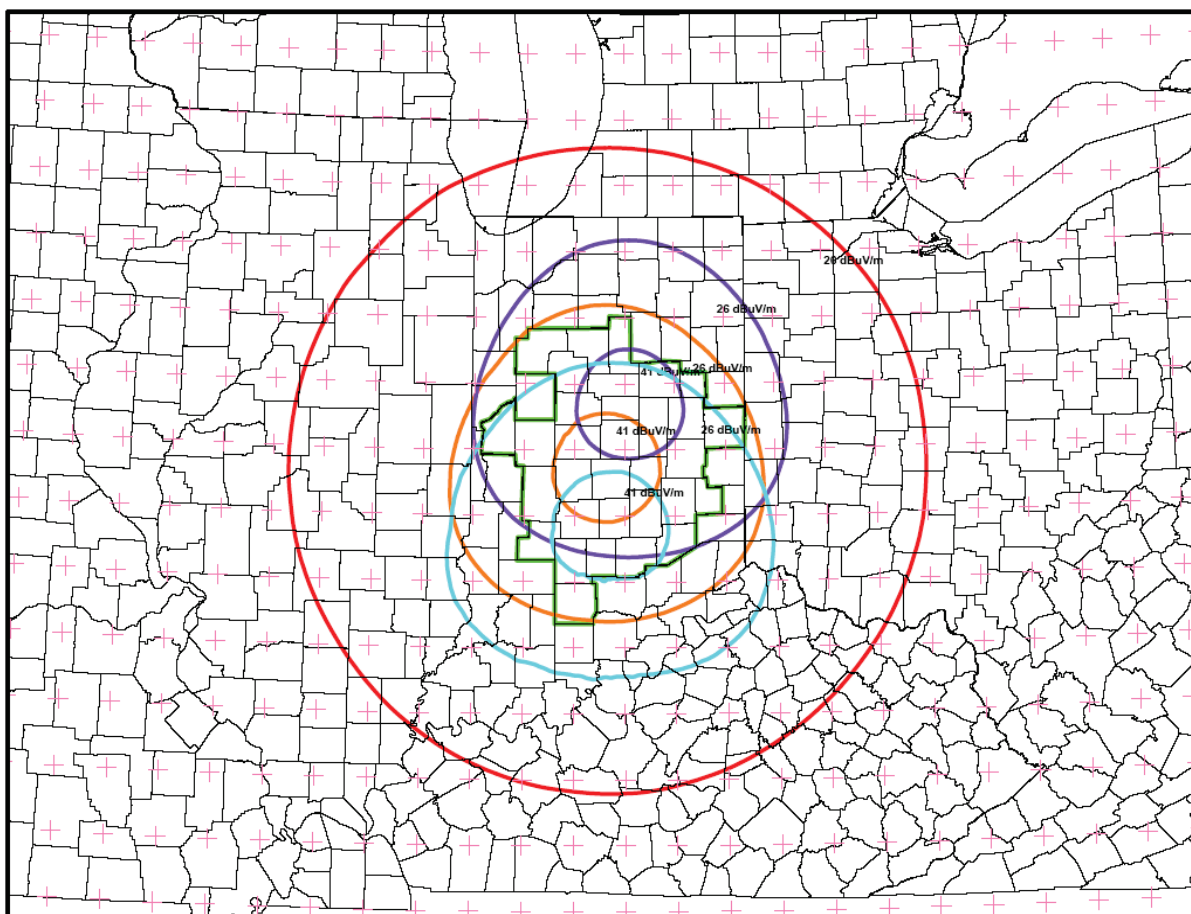


Figure A.8 Service and interference contours of three DTxTs of 50 kW at 75 meters vs interference contour of 1 MW transmitter at the same elevation.

Figure A.8 adds the interference contours of the three lower power transmitters of Figure A.7, and also superimposes the interference contour of the higher power transmitter at about the same elevation from Figure A.6. It can be seen that the interference contours of the group of three transmitters falls well inside the interference contour of the higher power facility.

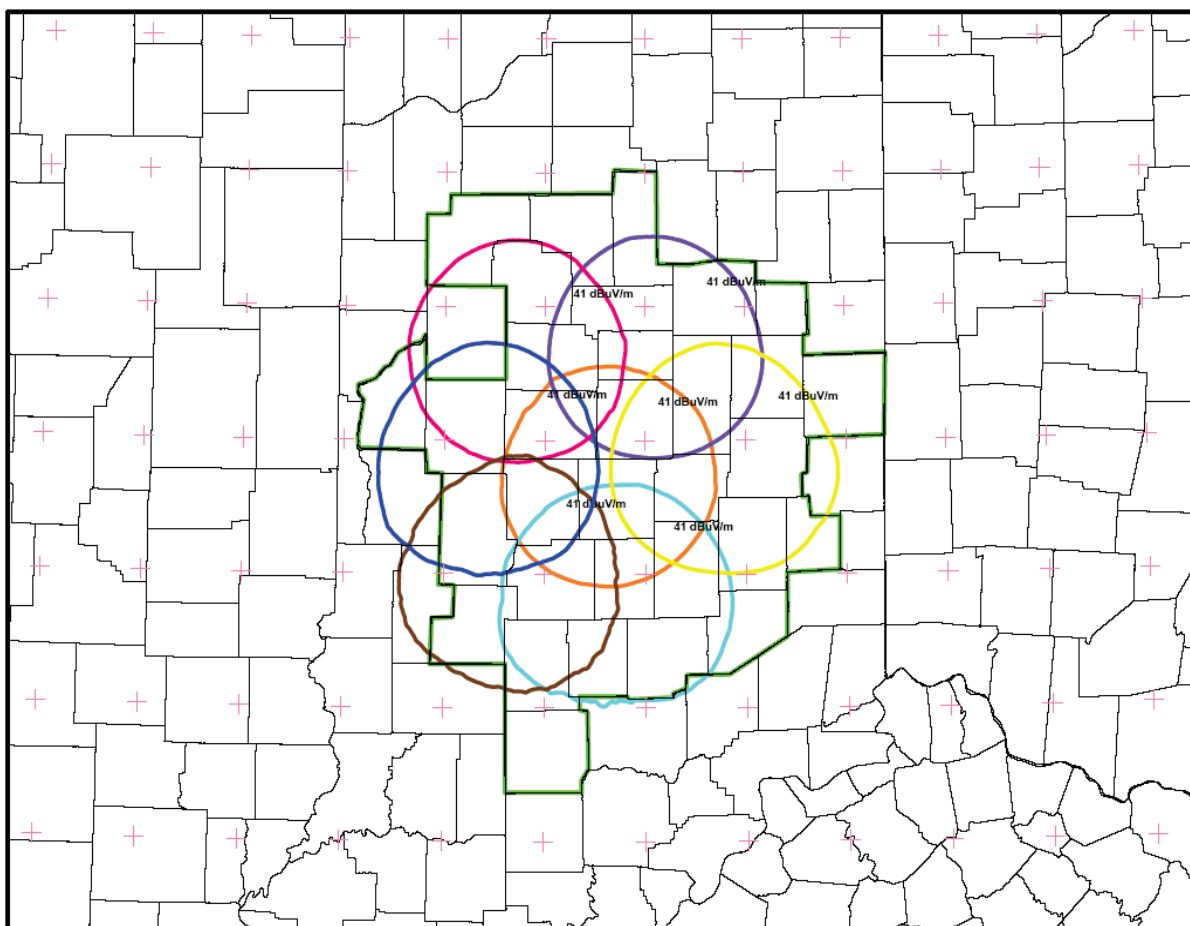


Figure A.9 Service contours of seven transmitters, each with 50 kW at about 75 meters above ground level.

Looking at what it might take to more completely cover the DMA with transmitters having intentionally overlapping coverage, Figure A.9 shows such a grouping. Unlike the first three, the transmitters added are arbitrarily located and have no relationship to existing towers. Should the locations selected for the creation of Figure A.9 prove to be optimal in a real implementation, then existing towers near those positions would be sought. Given the heights assumed for the various transmitters in this example, relatively tall cellular telephone towers near the target locations likely could be used for the purpose.

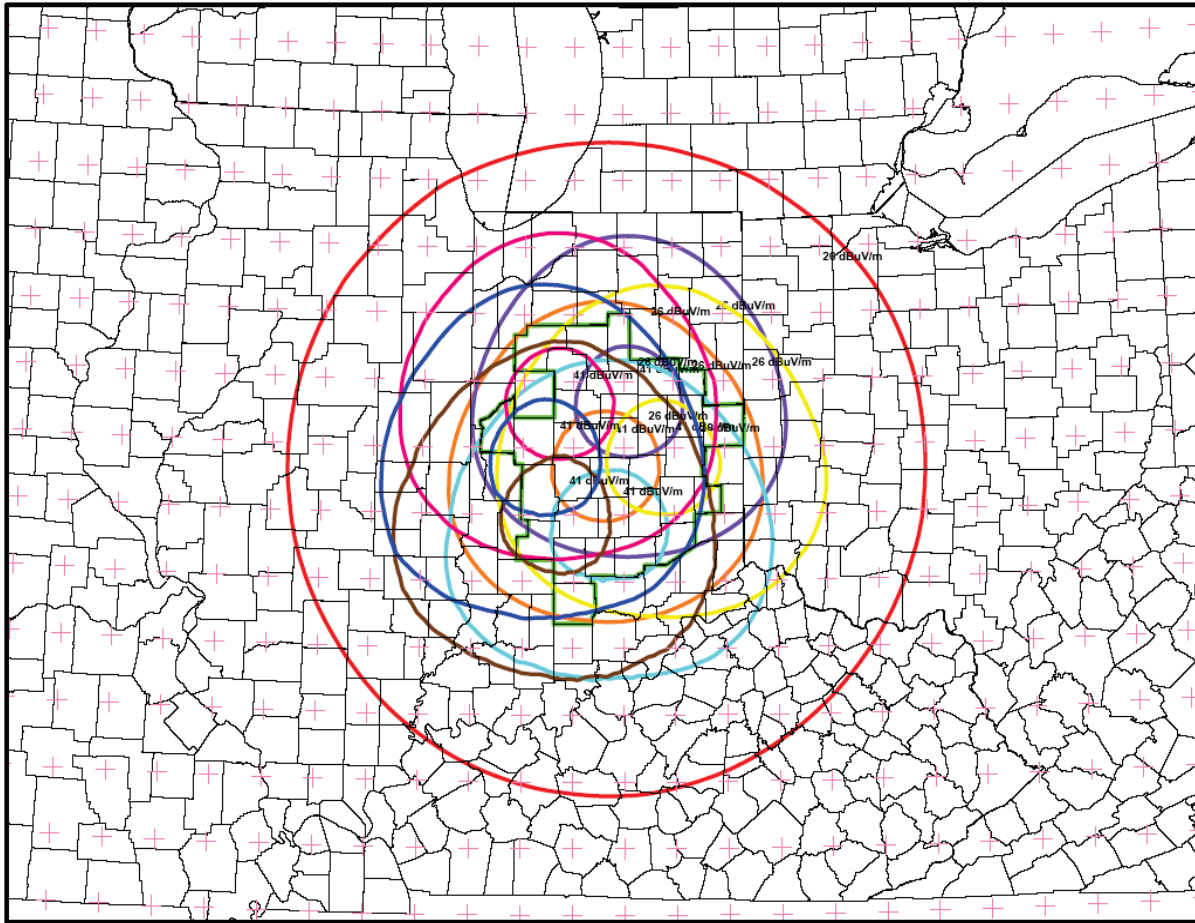


Figure A.10 Service and interference contours of seven DTxTs of 50 kW at 75 meters vs IX contour of 1 MW.

The interference contours of the same group of transmitters are shown in Figure A.10 in comparison to the interference contour of the higher power transmitter of Figure A.6. When, at each location on the map, the received power from all of the transmitters was aggregated, as it would be in a real interference study, it can be expected that the interference contour would extend a little beyond its current position. The amount of such an extension would be small, however, because the power increase in any direction would be no more than a few dB, with any corresponding contour extension therefore no more than a very few km.

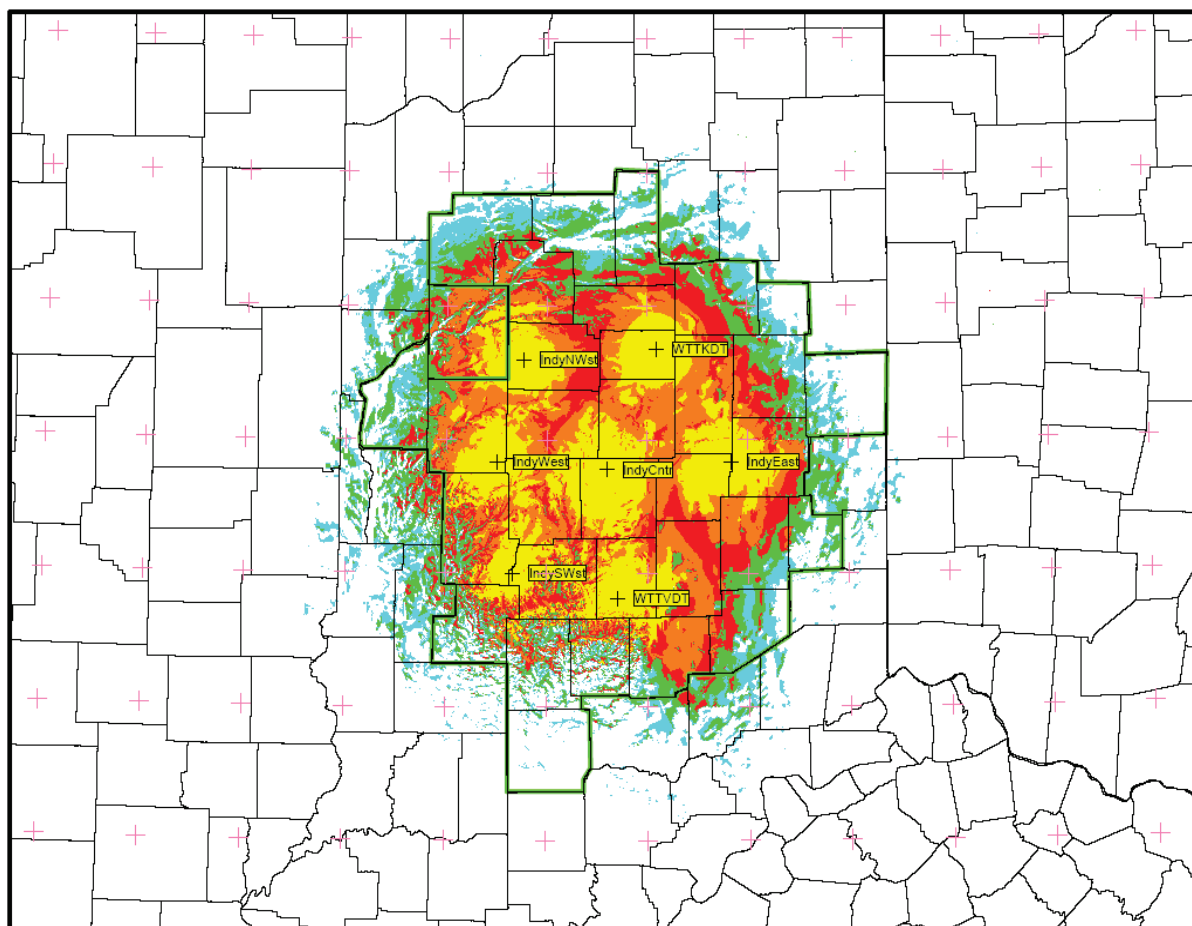


Figure A.11 Field strengths predicted from seven transmitters using Longley-Rice propagation model.

Returning to the type of performance that could be achieved from a system using a group of transmitters of the sort described, Figure A.11 shows a plot of the field strengths predicted at each location using the Longley-Rice propagation model. The locations of the transmitters that were indicated above as having been arbitrarily determined, in fact, were established to result in the sort of presentation shown in the Figure A.11 map. The objective was to provide a region in which the signals from contiguous transmitter service areas did not drop below the 60 dBu field strength value that is believed to be the minimum necessary for indoor antenna reception.

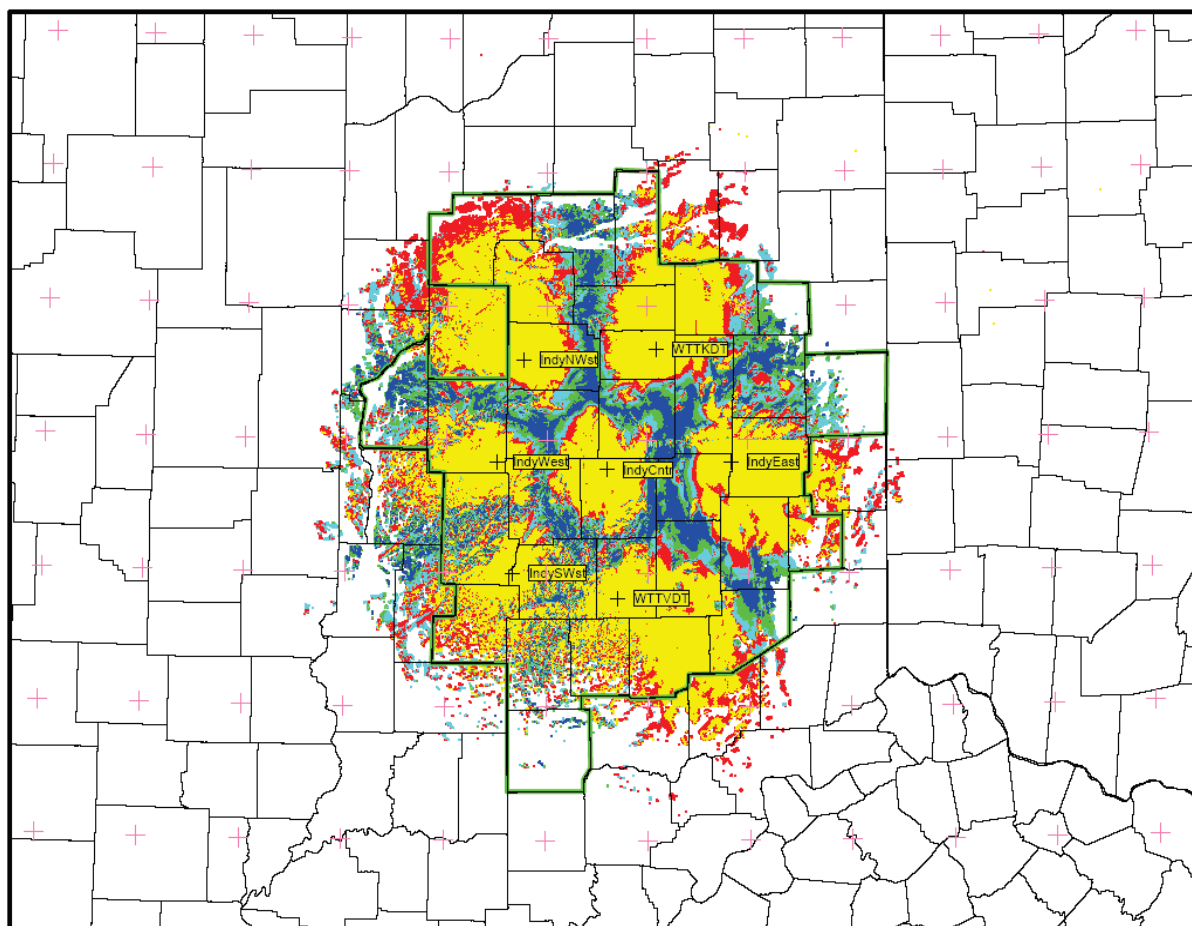


Figure A.12 Map of $C/(I+N)$ for seven DTxTs, each with 50 kW at 75 meters above ground level.

Figure A.12 shows a map of the $C/(I+N)$ values that would be used in managing the network internal interference between the transmitters in the group. Note that there are relatively well defined areas in which the signal from one transmitter is dominant (color coded yellow), and there are well defined areas in which the $C/(I+N)$ ratios fall below 20 dB (color coded other than yellow) and in which the network internal interference must be managed. The areas with the lowest $C/(I+N)$ ratios (in dark blue) tend to follow straight lines between the various pairs of transmitters.

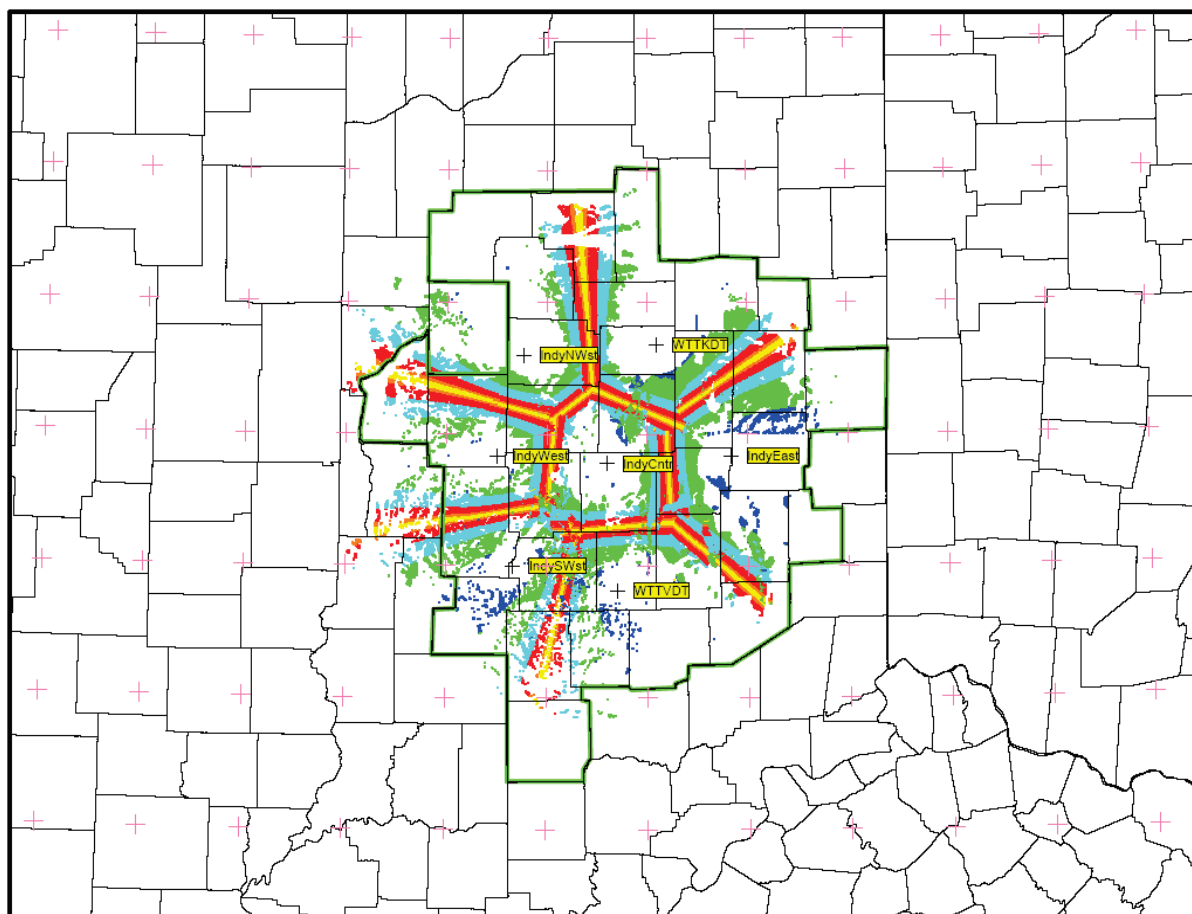


Figure A.13 Differential arrival times of signals from seven DTxTs, masked by 20 dB C/(I+N).

The relative times of arrival of the signals from the various transmitters are shown in Figure A.13. In this map, the arrival time information has been masked by the 20 dB C/(I+N) threshold so that only those areas with arrival time values relevant to the management exercise are presented. By comparison with Figure A.12, it can be seen that the areas with the lowest differential arrival times approximately superimpose over the areas having the lowest C/(I+N) values. This results in the elimination or near elimination of pre-echoes.

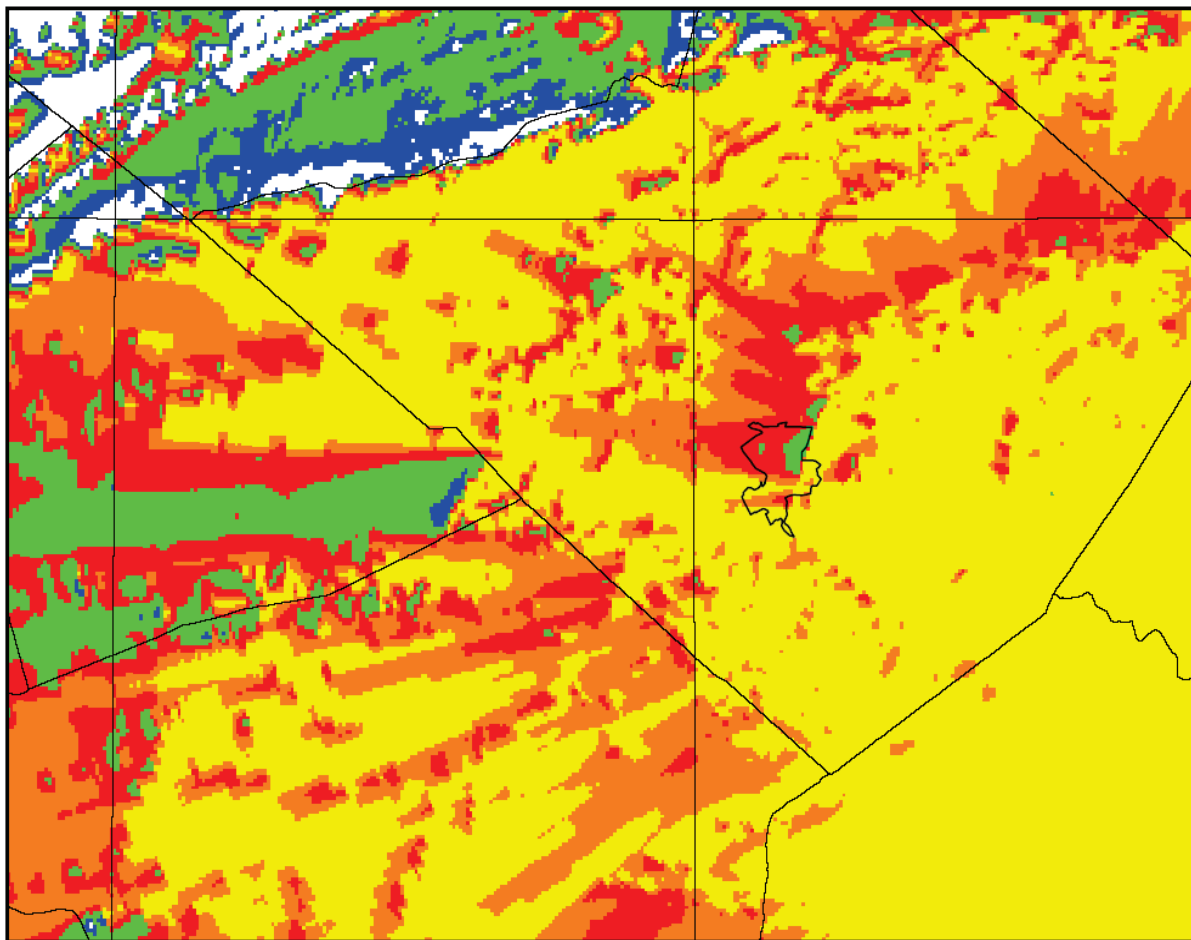


Figure A.14 Field strength map including city behind mountain with reduced signal levels (Longley-Rice).

A2.3 Example 3 — Increased Signal Levels in City Behind Mountain

Example 3 is a city in the Northeast that is behind and very close to a mountain that separates it from the transmitter site. As shown in Figure A.14, the field strength predicted with the Longley-Rice propagation model in many areas within the city are just above the noise limited threshold and not suitable for reliable indoor antenna reception. It is desired by the station involved to increase the field strength in the city to assure reliability of indoor antenna reception.

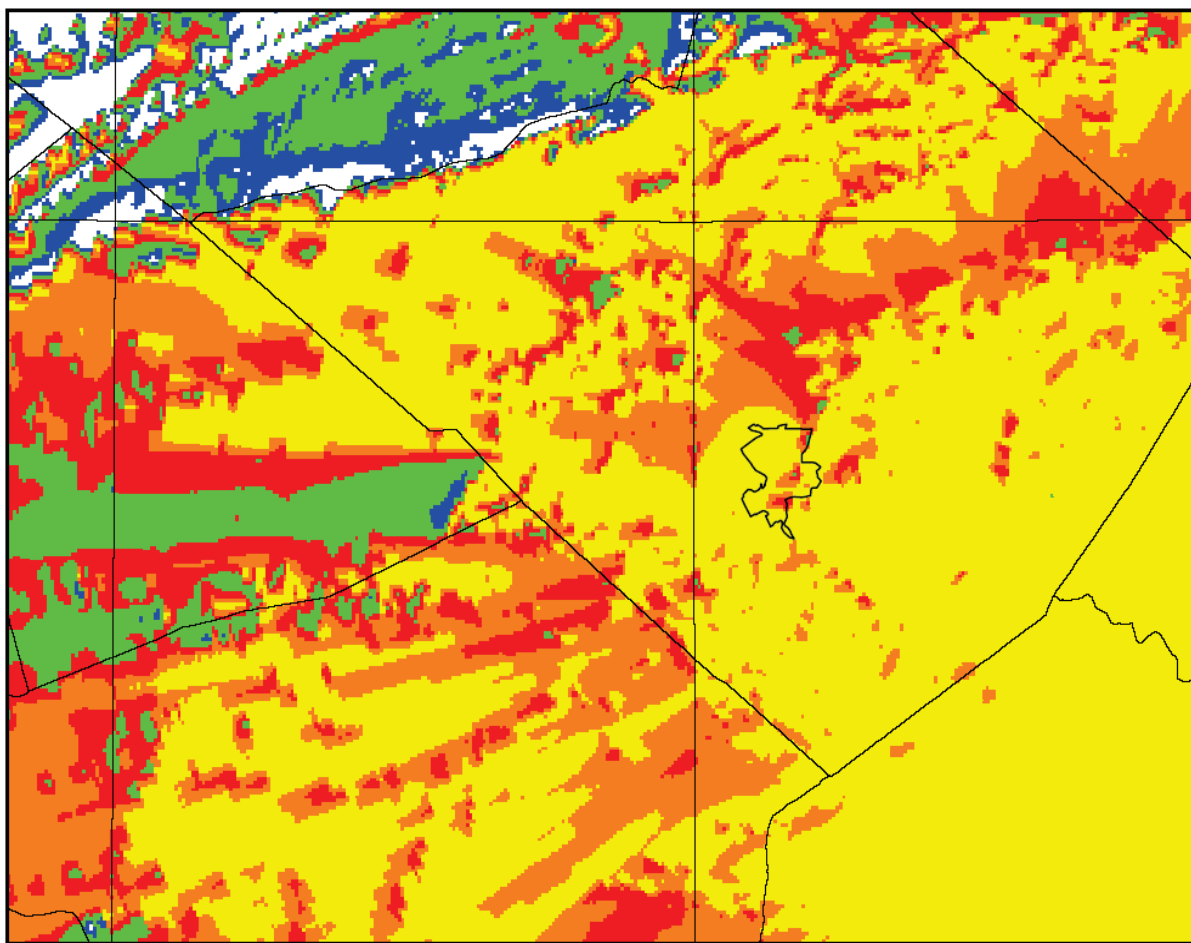


Figure A.15 City behind mountain with addition of DTxT showing more uniform signal levels.

Given the geography involved and the facilities already available, a Distributed Transmitter (DTxT) could be placed on the mountain that separates the city from the transmitter site. Alternatively, this could be a distributed translator (DTxR) or an EDOCR, but a DTxT is used in this example to permit demonstration of the effects of timing adjustments. Figure A.15 shows the field strength values when a DTxT of 1 kW ERP is added on the nearby mountain. Figure A.16 shows the $C/(I+N)$ resulting from the addition of the DTxT.

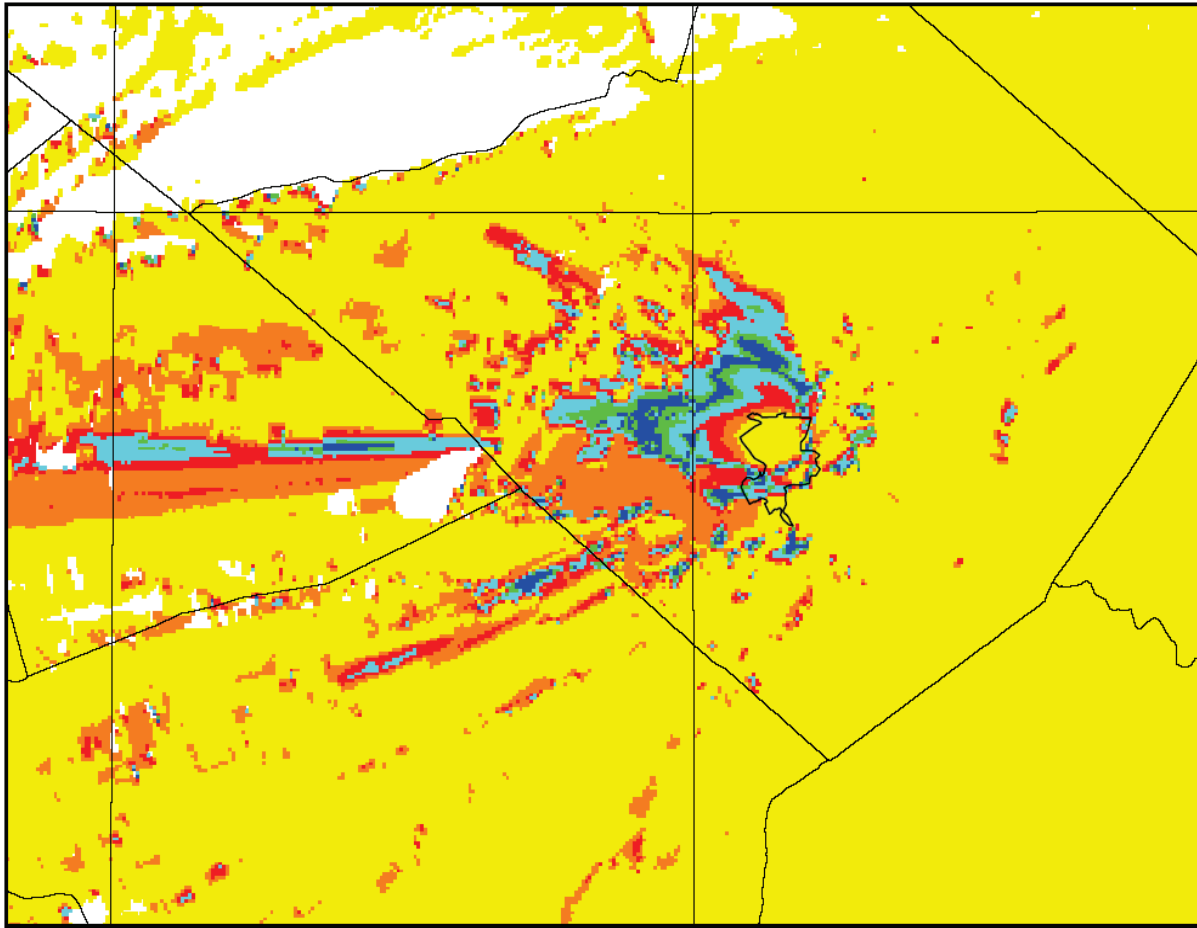


Figure A.16 $C/(I+N)$ map with addition of DTxT that makes signal levels in blocked city more uniform.

Figures A.17, A.18, and A.19 show the differential arrival times of the signals in the area served by the added DTxT and in the area served by the original transmitter in which the addition causes network internal interference. As before, the timing values shown are masked by a $C/(I+N)$ of 20 dB.

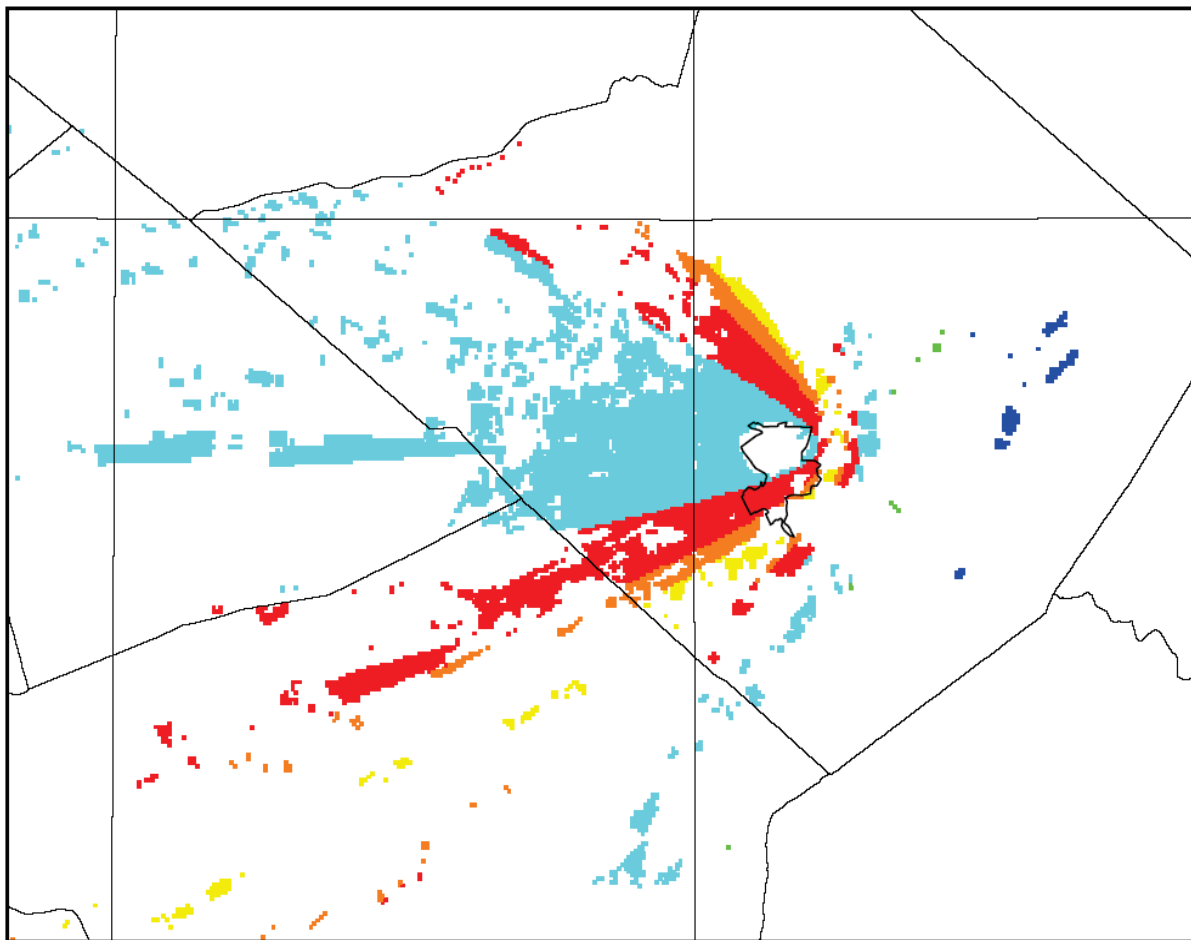


Figure A.17 Differential arrival times of signals from two transmitters with DTxT-2 delay offset = 45 μ s.

In Figure A.17, the emission times of the transmitted symbols of the added DTxT are delayed by 45 microseconds from the emission times of the symbols from the original transmitter. In Figure A.18, the emission times of the transmitted symbols of the added DTxT are delayed by 50 microseconds, and, in Figure A.19, the emission times are delayed by 55 microseconds. The differences in resulting delay spreads are readily apparent by comparing the three figures.

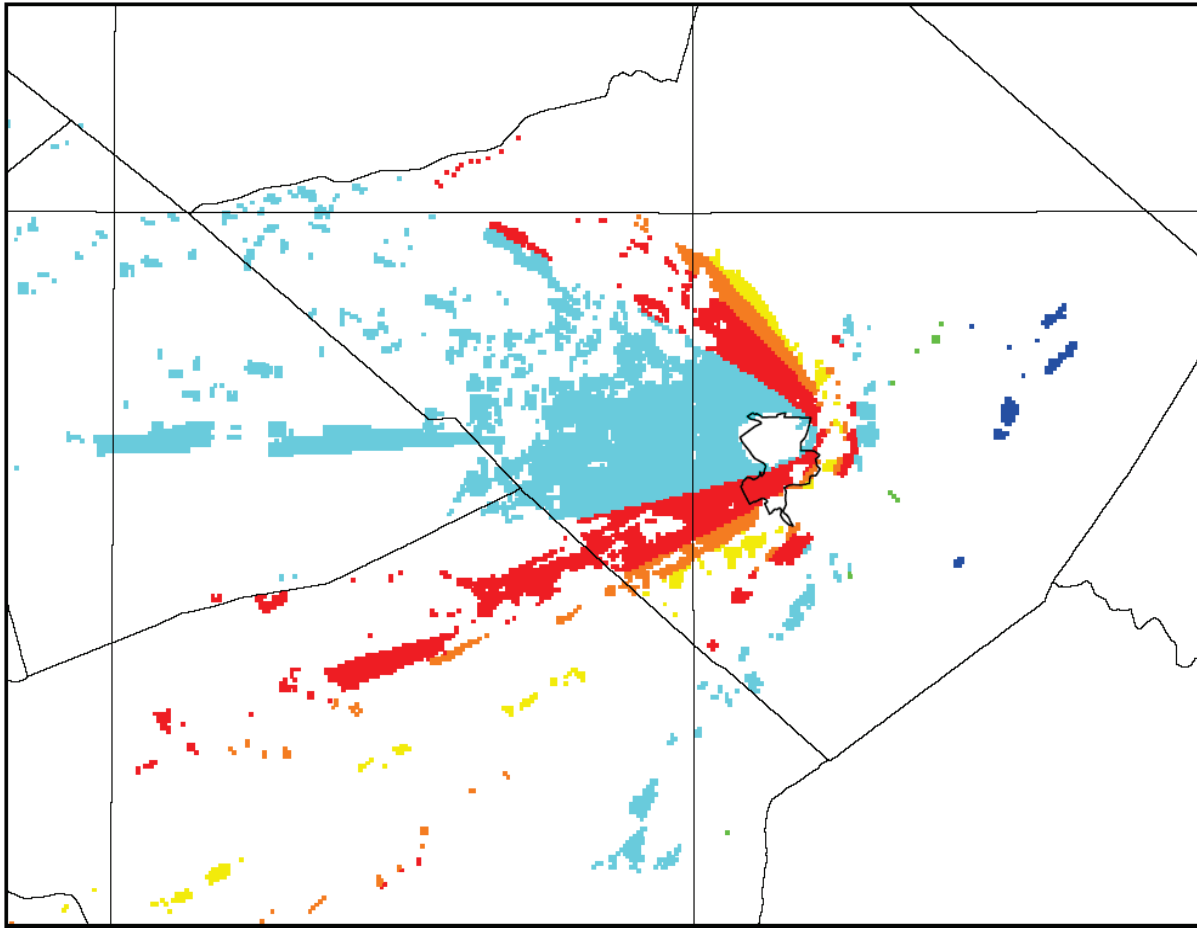


Figure A.18 Differential arrival times of signals from two transmitters with DTxT-2 delay offset = 50 μ s.

An adjustment approximating that in Figure A.18 is likely to be considered optimum from the standpoint of balancing differential arrival times across the geography involved in the interaction of the two transmitters. This conclusion results from the minimization of areas with delay spreads of 25 μ s and more (the green and dark blue areas) combined with delay spreads of 10 μ s or less throughout most of the area (as shown by the yellow, orange, and red areas). The choice of emission time delay offset would be confirmed best by analysis of populations in the areas receiving the signals within the various differential arrival time ranges.

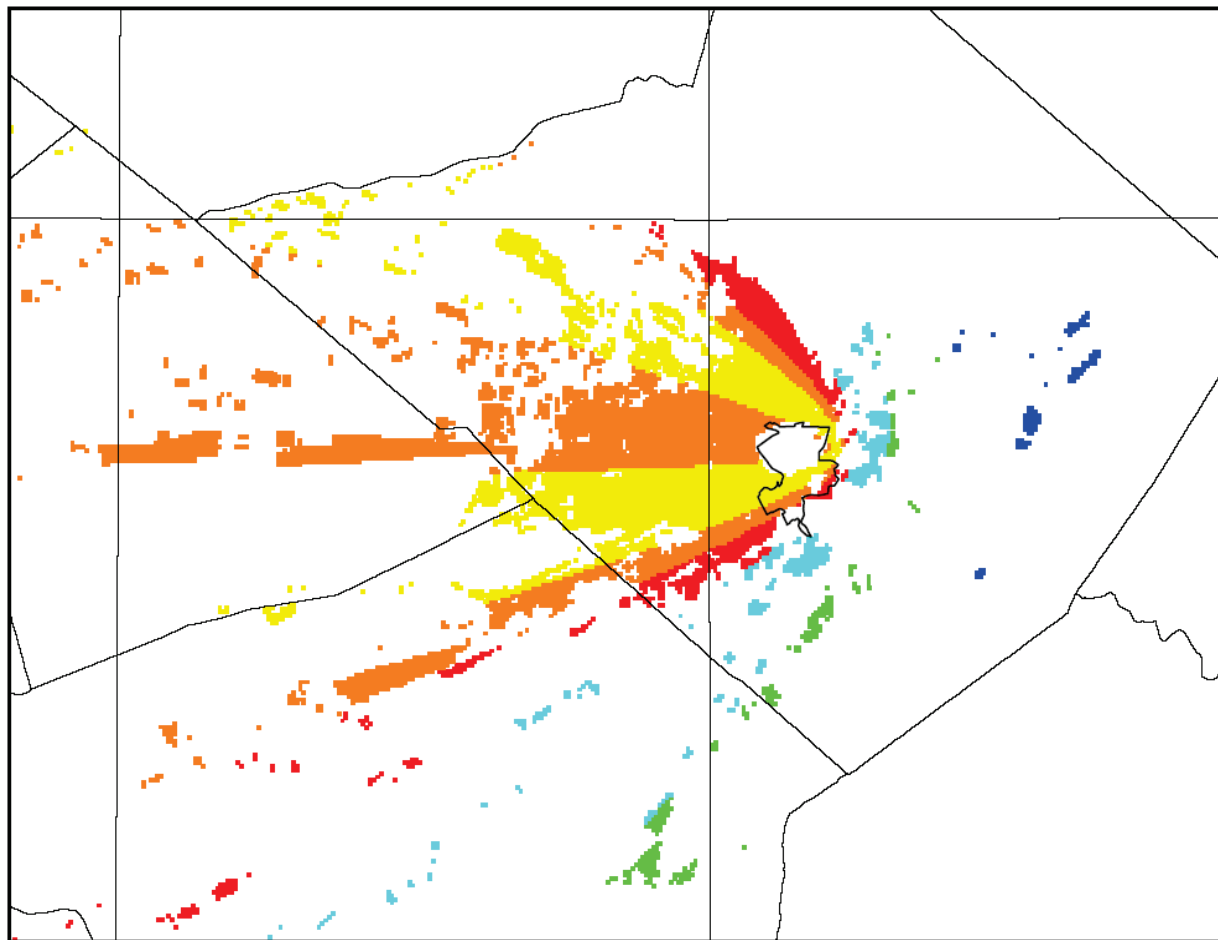
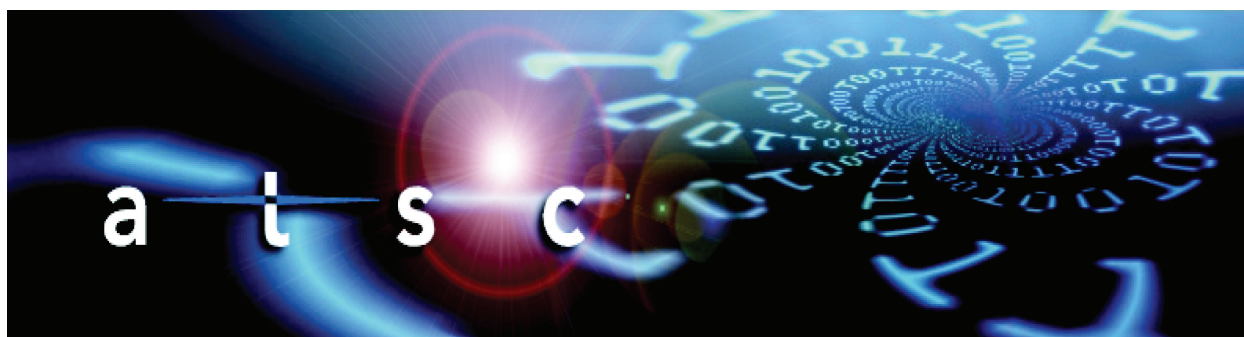


Figure A.19 Differential arrival times of signals from two transmitters with DTxT-2 delay offset = 55 μ s.



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