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An Evaluation of the Proposed Railroad VHF Band Channel Plan

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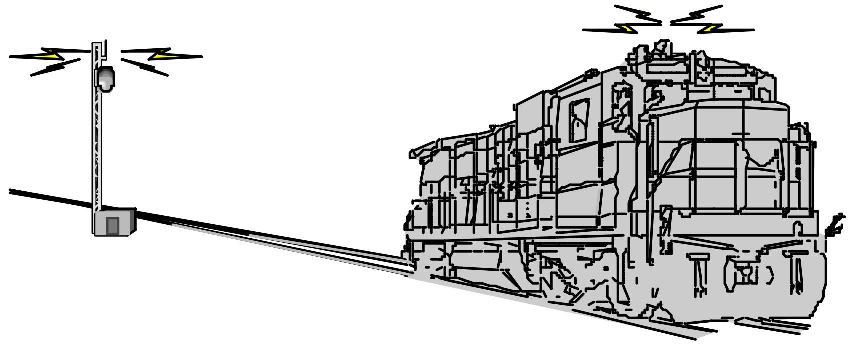
John M. Vanderau and Eldon J. Haakinson*



As a result of the Federal Communications Commission's radio spectrum realignment initiative, land mobile radio manufacturers are being encouraged to incorporate narrowband and trunking technologies into their designs in order to efficiently utilize the radio spectrum. As new equipment designs become available in the marketplace, users must infuse these new systems into their existing land mobile radio infrastructures. The railroad industry is developing a strategy to accomplish this migration. This strategy includes defining a band plan for the efficient use of the railroad VHF spectrum. In this report, the authors evaluate the railroad industry's rechannelization strategy, offer an alternate approach, and conclude with comments regarding the efficacy of both railroad VHF band plans.

Keywords: Band plan, channel plan, land mobile radio, WCTF, Project 25, railroad, spectrum realignment, TIA-102, VHF

1 INTRODUCTION

The Federal Communications Commission (FCC) has encouraged land mobile radio (LMR) manufacturers to incorporate narrowband (12.5-kHz bandwidth or smaller) and trunking technologies into their designs to more efficiently utilize the radio spectrum, particularly in the crowded very high frequency (VHF) bands [1]. Failure to comply with this directive could result in the denial of FCC equipment type acceptance. Whereas this directive does not affect current equipment designs, as the state-of-the-art in radio technology continues to advance, new radio designs will incorporate digital narrowband

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technologies. Users should begin planning now to deal with migration and compatibility issues when they introduce new equipment and systems into their existing LMR infrastructures.

To address this issue within the railroad community, the railroads, radio equipment manufacturers, and Government formed the Wireless Communications Task Force (WCTF), an ad-hoc committee dedicated to solving radio communications issues unique to the North American railroad industry. WCTF is currently considering how to best migrate the railroad industry's existing 160-MHz wideband (15-kHz bandwidth) analog LMR equipment to more modern, spectrally-efficient narrowband digital equipment.

In order to utilize the railroad VHF spectrum more efficiently, WCTF has proposed a channel assignment scheme that accommodates trunking technology solutions and uses the narrower channel spacings of the realigned railroad VHF spectrum. WCTF has adopted radio equipment conforming to the Telecommunications Industry Association (TIA) land-mobile radio standard TIA-102 [2], also known as Project 25.

The Federal Railroad Administration (FRA) wants to ensure that if the railroads adopt this VHF band plan and the TIA-102 standard, railroad communications will not be degraded. The Institute for Telecommunication Sciences (ITS) was tasked by the FRA to determine whether the proposed rechannelization plan will provide the radio coverage and system capacity needed to achieve these objectives.

2 SYSTEM DESIGN ISSUES

In order to assess the viability of the proposed band plan, several issues must be investigated. These issues are:

- Performance of new equipment — within the constraints of the given channel assignments, will the new equipment perform adequately? Is it backward-compatible with legacy analog FM systems now in service? Is the new equipment a proprietary design, or does it conform to recognized industry standards, thereby ensuring interoperability between different vendors' equipment?

- Channel assignments — are there enough radio channels to support all the business processes of the railroad industry? Is the voice traffic loading equally distributed across the available channels? Are there any techniques that can be employed to increase the traffic capacity?

- Interference issues — will the proposed new equipment cause interference to the older existing equipment? Will the older systems cause interference to the new systems? If repeaters and base stations will be collocated (such as would be the case for trunked systems), what is the likelihood that intermodulation interference will be present? If present, can it be adequately suppressed? Are receivers likely to be subject to desensitization? Will cochannel and adjacent-channel interference be a problem?

- Coverage — For a given level of speech intelligibility or effective data throughput, will the proposed radio system provide adequate geographic coverage?

The authors explore these topics, evaluate options, and provide procedural guidelines that LMR systems planners can follow when designing a new system. The proposed railroad band plan is described in Chapter 3. In Chapters 4 through 8, the impact of equipment selection, collocation of repeaters and intermodulation interference, VHF propagation, adjacent-channel and cochannel interference, cellular topologies, and trunking on the proposed band plan is described. Chapter 9 presents an alternate plan and compares it to the railroad's plan.

3 RAILROAD VHF BAND PLAN

The existing railroad VHF band plan consists of 181 frequencies, spaced every 7.5 kHz, from 160.215 to 161.565 MHz [3]. The band plan proposed by WCTF is shown in Figure 1. The two blocks of spectrum from 160.215 to 160.8075 MHz and 160.935 to 161.5275 MHz comprise 80 trunked duplex voice channel-pairs. The two blocks of spectrum comprised of 5 channels apiece (from 160.815 to 160.845 MHz and 161.535 to 161.565 MHz) are non-trunked duplex channel-pairs. The remaining block of spectrum consists of 11 simplex channels (from 160.8525 to 160.9275 MHz).

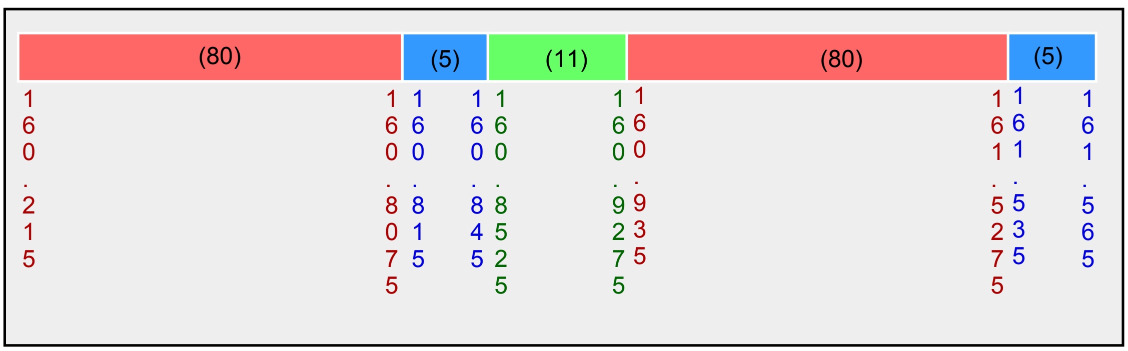

Figure 1. Railroad VHF band and WCTF's band plan.

Trunking permits many users to share a finite number of radio channels, similar to how a finite number of telephone lines interconnecting telephone central office switches can be shared by a large number of individual telephone subscribers. These eighty duplex channel-pairs are apportioned among ten trunk groups. Each trunk group has eight channel-pairs; each channel-pair uses one frequency (channel) for transmitting and a different frequency (channel) for receiving. At each radio base station site, collocated transmitters and receivers are assigned operating frequencies from a given trunk group. Frequency assignments at an individual site are not commingled with frequencies from other trunk groups. The trunk groups are interstitially placed in the frequency domain, as depicted in Figure 2.

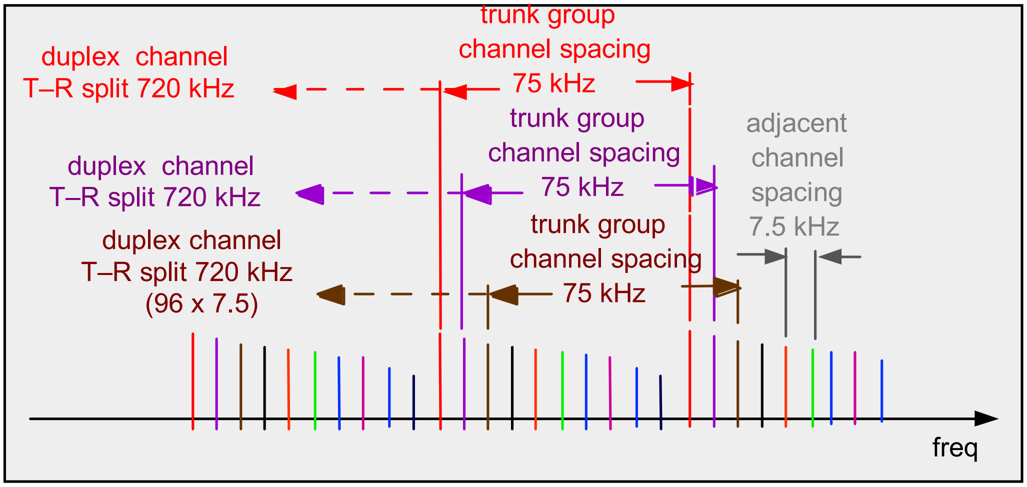

Figure 2. WCTF-proposed channel spacing. Trunk groups are interstitially placed in the frequency domain.

All frequencies belonging to a particular trunk group are depicted with the same color (and same vertical line length). Different trunk groups are differentiated by different colors (and different line lengths). Within a given trunk group, adjacent channels are spaced 75 kHz apart, and the frequency split between transmit and receive frequencies of any particular channel-pair is 720 kHz. The 75-kHz channel spacing within any particular trunk group allows all ten trunk groups to be "interlaced" in the frequency domain at 7.5-kHz channel spacings.

4 EQUIPMENT STANDARD

WCTF adopted TIA-102 as the radio standard for railroad radio systems in February 1996. ITS agrees that adoption of this standard has merit because:

- TIA-102 provides a migration path by allowing existing legacy analog FM radio equipment to interoperate with the newer, dual-mode narrowband, trunking-capable equipment,
- adoption of the TIA-102 standard promotes interoperability between different user communities, which could provide significant benefit in mutual aid situations between the railroads, public safety, and other critical infrastructure services, and
- TIA-102-compliant equipment is capable of voice-only, data-only, or voice-plus-data operation, thereby eliminating logistics requirements for separate radios to carry out these desired functions.

First-generation TIA-102 radios have a 12.5-kHz bandwidth, use digital modulation, and are backward-compatible with analog frequency-modulated (FM) equipment, supporting both 25- and 12.5-kHz bandwidths. Minimum performance parameters are cited in various TIA documents [2, 4]. Second-generation TIA-102 radios will utilize 6.25-kHz bandwidths, but presently the characteristics for these radios are not well established.

4.1 Receiver Sensitivity

ITS measured the digital and narrowband analog receiver sensitivity of a typical TIA-102 radio [5] by following TIA's land mobile radio and digital transceiver performance measurements documents [6, 7]. This data is presented in Figure 3. Measured sensitivities were better than the manufacturer-advertised typical sensitivities for this radio in both modulation formats.

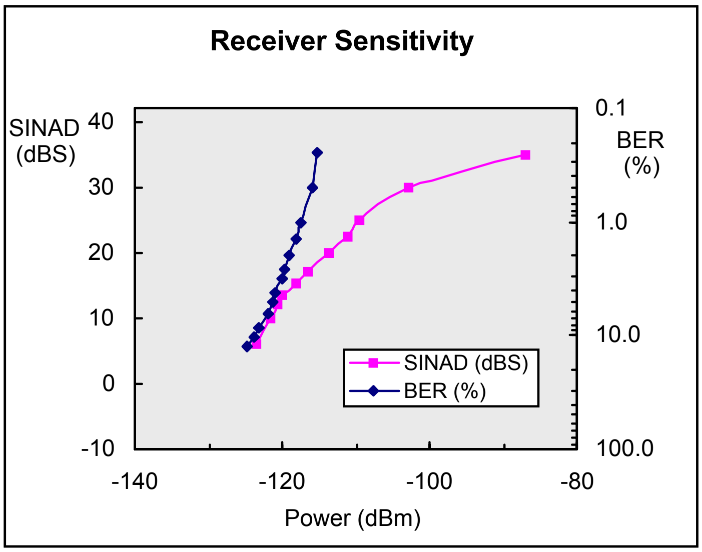

Figure 3. Measured receiver sensitivity of a typical TIA-102 radio.*

Delivered audio quality (DAQ) is a figure of merit that indicates the intelligibility of a speech signal,† and ranges in value from "1" to "5" [4]. ITS estimated the DAQs of demodulated speech signals by using an automated computer technique that computed the

* Sensitivity for analog FM modulation format was measured using 2.5-kHz frequency deviation. For wider bandwidths, sensitivity can improve by as much as $20 \log(\Delta f_{wb} / \Delta f_{nb})$, where Δf_{wb} and Δf_{nb} are the wideband and narrowband (2.5-kHz) frequency deviations.

† Delivered Audio Quality Metrics:

DAQ 1	Unusable. Speech present but not understandable.
DAQ 2	Speech understandable with considerable effort. Requires frequent repetition due to noise/distortion.
DAQ 3	Speech understandable with slight effort. Requires occasional repetition due to noise/distortion.
DAQ 3.4	Speech understandable without repetition. Some noise/distortion present.
DAQ 4	Speech easily understood. Occasional noise/distortion present.
DAQ 5	Speech easily understood.

auditory distance[*] between the received speech signal and its originally transmitted (reference) speech signal [8]. This was done at a number of SINAD and BER reference sensitivities. Auditory distance was then mapped to a DAQ score by using a previously determined empirical relation. Figure 4 shows the DAQ values of speech signals, as determined by ITS, at power levels corresponding to various BERs and SINAD ratios. Also shown in the figure are TIA's reference DAQ values.

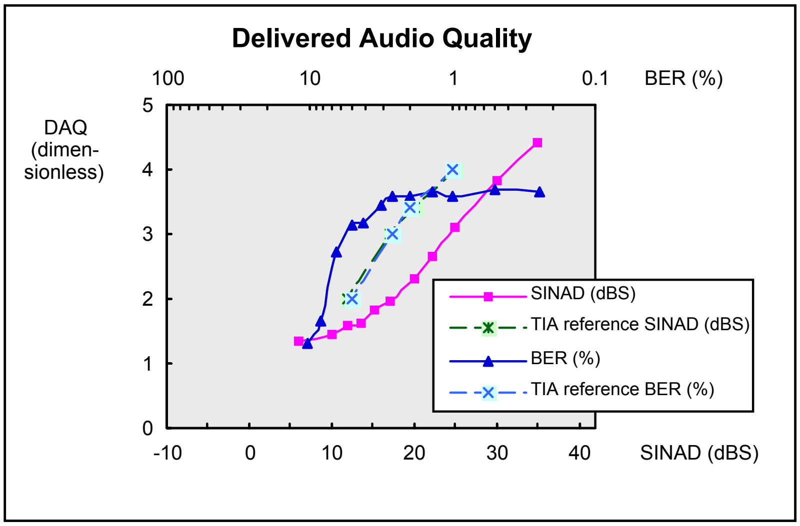

Figure 4. DAQ vs. SINAD and BER.

A threshold effect is apparent in the ITS digital DAQ plot, above which DAQ values remain relatively constant versus BER. This indicates an upper limit on the audio quality from analog-to-digital conversion. It also demonstrates the error-correcting capability of the TIA-102 vocoder. At operating points below this threshold (greater BER), the DAQ begins to deteriorate rapidly. But even in this region, the perceived speech quality of the digital signal still exceeds that of an analog signal.

This information can be used when determining the geographic coverage. The greater the minimum level of speech intelligibility required, the smaller the geographic coverage area will be for "acceptable" speech quality. Chapter 6 provides example computations to determine the radio coverage area.

4.2 Cochannel and Adjacent-Channel Rejection

TIA-102 performance specifications [2] assume a channel spacing of 12.5 kHz or more. However, since the railroad VHF band's channel spacing is only 7.5 kHz, spectrum

[*] Auditory distance measures the perceived difference between two speech signals. When one speech signal is a "perfect" reference signal, and the other speech signal is a distorted signal (e.g., radio output), auditory distance corresponds to the perceived quality of that distorted signal relative to the reference signal. When two speech signals are identical, the auditory distance between them is zero. As the perceived differences between the two signals increase, the auditory distance increases.

overlap will occur between signals on adjacent channels. Therefore, adjacent-channel rejection performance in 7.5-kHz channel-spacing environments is of interest.

Under varying conditions of adjacent-channel and cochannel interference, ITS measured the DAQs of a typical TIA-102 radio's demodulated speech signals [5], in a manner like that discussed in Section 4.1. Results are presented in Figure 5.

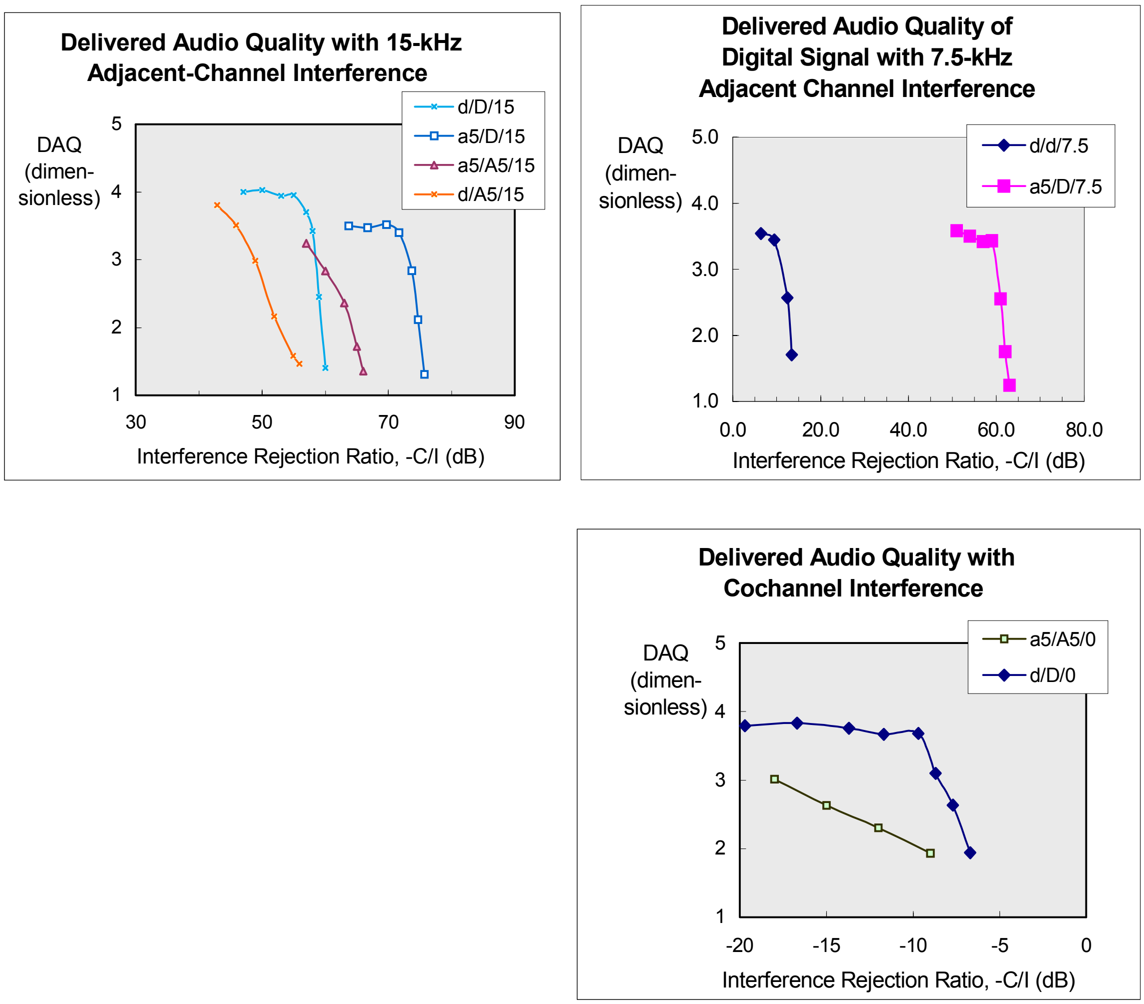

Figure 5. *Delivered audio quality* of demodulated speech signals under various interference conditions.

The *interference modes* depicted in the plots are:

(1) a digital signal interfering with a desired digital channel (d/D),
(2) a wideband[*] analog signal interfering with a desired digital channel (a5/D),
(3) a digital signal interfering with a desired wideband[*] analog channel (d/A5), and
(4) a wideband[*] analog signal interfering with a desired wideband[*] analog channel (a5/A5).

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[*] 5-kHz peak frequency deviation.

Adjacent-channel and cochannel rejection performance impacts the physical separation distance requirements between a receiver and undesired transmitters. The greater the level of speech intelligibility required, the greater the distance between the receiver and unwanted transmitters must be. This, in turn, influences base station siting and frequency reuse planning parameters. In Chapter 6, a sample procedure is presented that delineates the necessary steps to determine the required geographical separation between a receiver and interfering adjacent-channel and cochannel transmitters.

5 CO-SITE ANALYSIS

This chapter reviews some of the methods used that allow multiple transmitters and/or multiple receivers to share a common antenna at a fixed repeater or base station site. Following that, example procedures are given to:

- identify possible receiver intermodulation frequencies,
- estimate power levels of intermodulation interference, and
- determine the degradation in noise figure and intermodulation rejection performance.

5.1 Transmitter Combining Techniques

There are several methods to combine multiple transmitters so they can share a common transmitting antenna. Typically, hybrid ferrite combiners or cavity combiners are employed. These devices can combine RF signals at moderate power levels.

5.1.1 Hybrid Ferrite Combiners

Hybrid ferrite combiners, commonly referred to as hybrids, are useful in cases where frequency separation between the combined transmitters is very small. They provide very good isolation between transmitters closely spaced in frequency, are compact, and are easily expandable to accommodate additional transmitters.

Hybrids are, however, more lossy than the bandpass-isolator type of combiners. A hybrid combines two signals into one composite signal. Its insertion loss is approximately 3.2 dB per hybrid; that is, a hybrid will reduce the output power of each signal presented to it by about 3.2 dB. Figure 6 illustrates how hybrids can be used to combine up to four RF signals.

Losses would be about 6.4 dB per channel (because there are two "tiers," or "layers," of hybrid assemblies, the insertion loss is $2 \times 3.2 = 6.4$ dB). Extending this approach to combine up to eight channels would require three "tiers" of hybrids and isolators, yielding a per-channel insertion loss of about 9.6 dB. The limitations of using hybrids to combine multiple transmitters quickly becomes apparent, with each factor-of-two increase in the total number of combined transmitters adding 3.2 dB of attenuation to each channel.

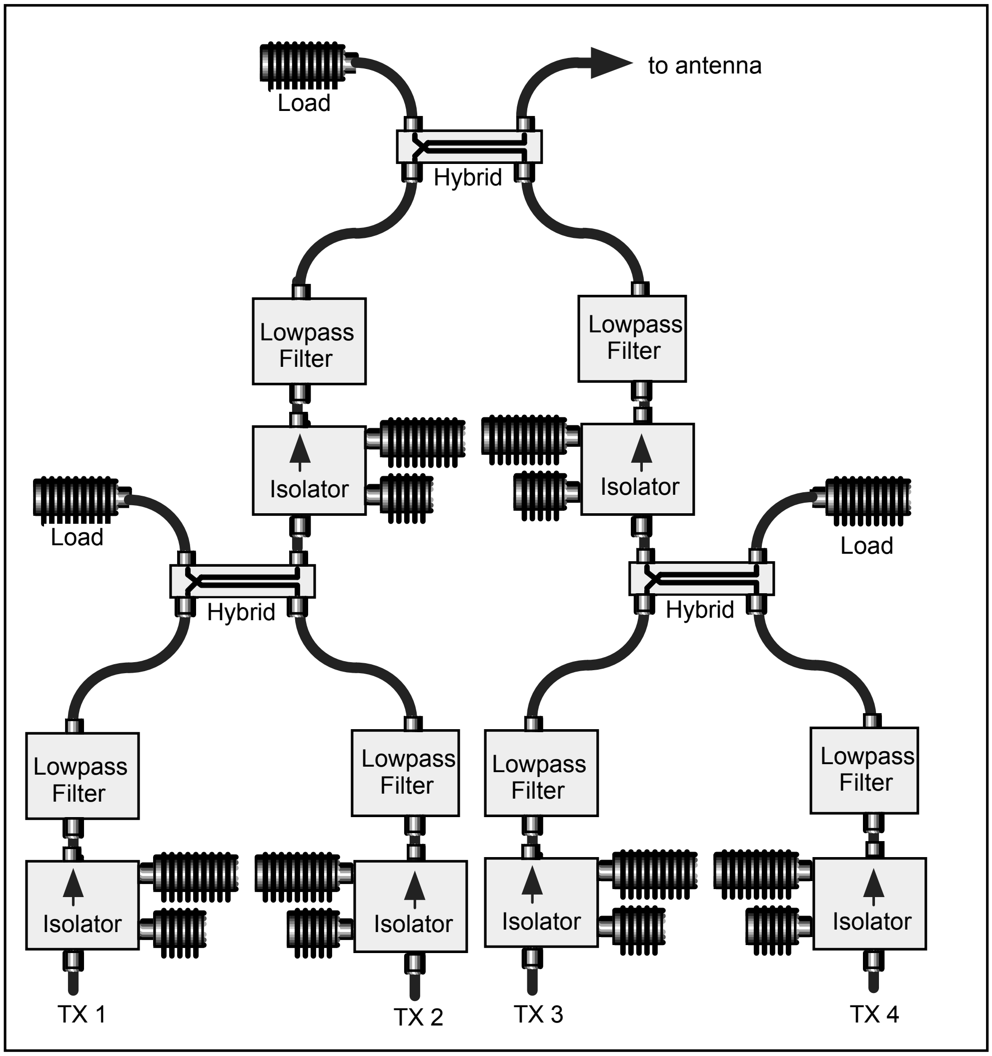

Figure 6. Four-channel transmitter hybrid-ferrite combiner.

5.1.2 Bandpass-Isolator Combiner

Bandpass-isolator combiners are constructed with ferrite isolators and band-pass cavities. One such design is shown in Figure 7. Adjacent-channel rejection and spurious signal suppression can be improved by cascading additional cavities in each transmitter's signal path. A cavity's insertion loss is about 0.5 to 3 dB at its resonant frequency. The insertion loss depends on the cavity's selectivity (bandwidth and "steepness," or "roll-off" of the passband edges). Adjustable coupling loops determine a cavity's selectivity. Insertion loss increases as the coupling loops are adjusted to provide more selectivity.

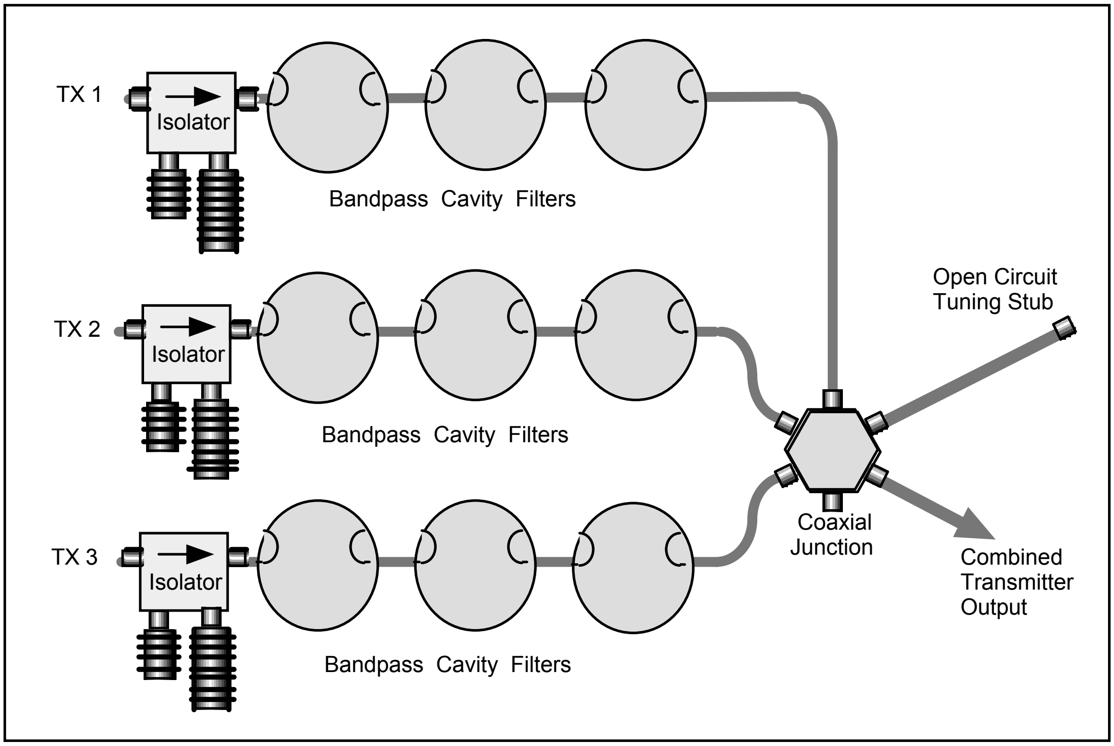

Figure 7. Bandpass-isolator combiner.

Channel spacings cannot be as close in frequency as they can with hybrid ferrite combiners. Most manufacturer product sheets recommend channel spacings on the order of 200 kHz or so, but closer channel spacings are possible. Closer spacing requires cavities to be tuned for the narrowest passband response (highest insertion loss), or a greater number of cavities (each having less overall insertion loss, but wider frequency passband) to be cascaded in series (greater physical volume and cost), or both.

5.1.3 Combination of Hybrid- and Cavity-Combining Solutions

Figure 8 illustrates a composite approach that utilizes both hybrids and cavities to overcome the drawbacks of using only hybrids (high insertion loss for eight channels) or only cavities (limited selectivity for close channel spacings).

In this scheme, “odd” channels, spaced every 150 kHz, are combined using cavities. “Even” channels, also spaced every 150 kHz, are combined in a similar fashion. Finally, these two “even” and “odd” channel banks are interstitially combined using a hybrid coupler. The two channel banks are offset in frequency from one another by 75 kHz.

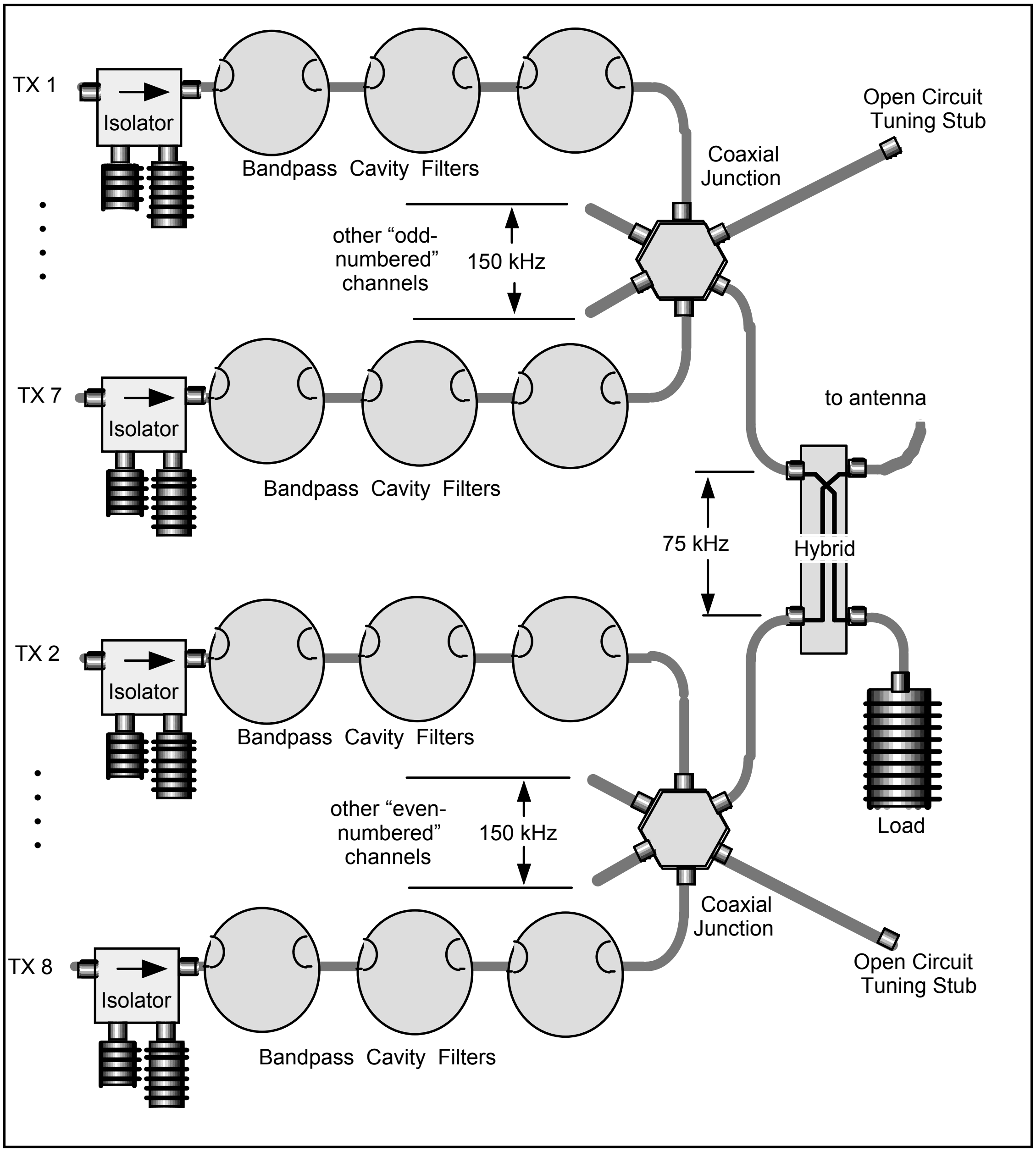

Figure 8. Composite cavity/hybrid combiner.

5.2 Receiver Multicouplers

To combine multiple independent radio receivers to a common receiving antenna system, multicouplers are employed. Receiver multicouplers divide a multichannel signal into several discrete receiver channels. This division reduces the strength of each signal delivered to each individual receiver.

Preceding the multicoupler's RF splitter with a preamplifier will offset these losses. If the gain of the preamplifier is too great, though, the preamplifier will cause unacceptable levels of intermodulation interference to the receivers that follow. Similarly, if the gain is too low, receiver sensitivity will be worse than for an isolated receiver with no multicoupler. Section 5.5 gives mathematical formulas to determine the effects of multicoupler preamplifier gain on system noise figure, system sensitivity, and intermodulation distortion.

5.3 Duplexers

Duplexers permit simultaneous operation of transmitters and receivers with a single antenna. They are constructed from parallel sets of bandpass/band-reject filters. They route RF energy flow from a transmitter to an antenna at one frequency, and from an antenna to a receiver at a different frequency, while preventing the flow of RF energy directly between transmitter and receiver. Figure 9 shows a block diagram of a cavity-type duplexer.

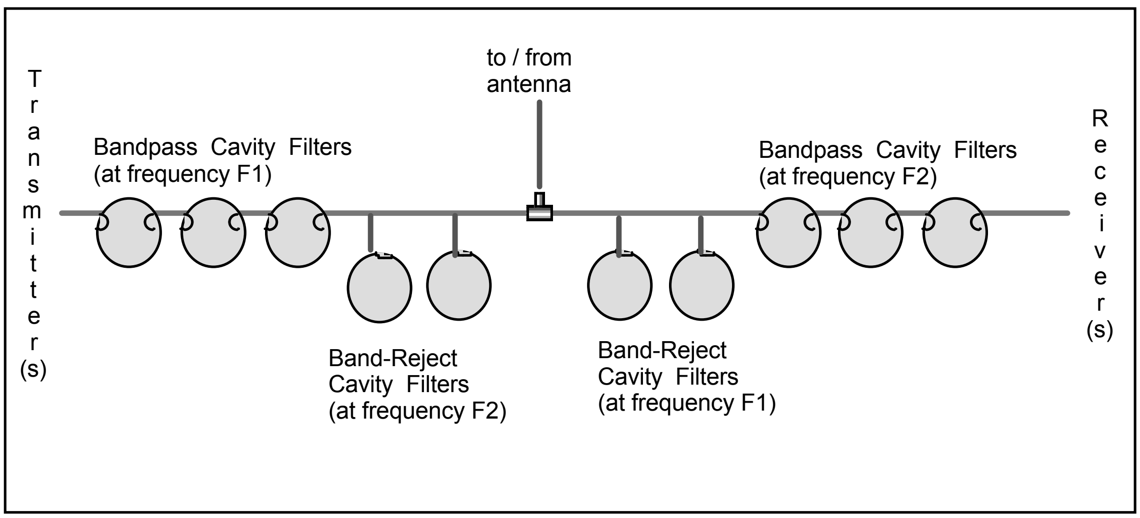

Figure 9. A duplexer constructed from parallel cavity filters.

Figure 10 shows a typical frequency response of a duplexer that has a 600-kHz separation between transmit and receive frequencies.* For wider frequency splits (such as 720 kHz), the curves are shaped somewhat differently, but the same general behavioral characteristics will hold. These characteristics include a sharp "notch" (large amount of attenuation) around some desired rejection frequency, minimum attenuation around some

* The data shown here are representative of manufacturers' product sheets for a 600-kHz frequency split at VHF. Of the publicly available product data reviewed, this frequency split was the closest approximation to the proposed band plan's 720-kHz frequency split.

desired pass frequency, and "moderate" attenuation of signals at frequencies more distantly removed from the resonant pass and notch frequencies.

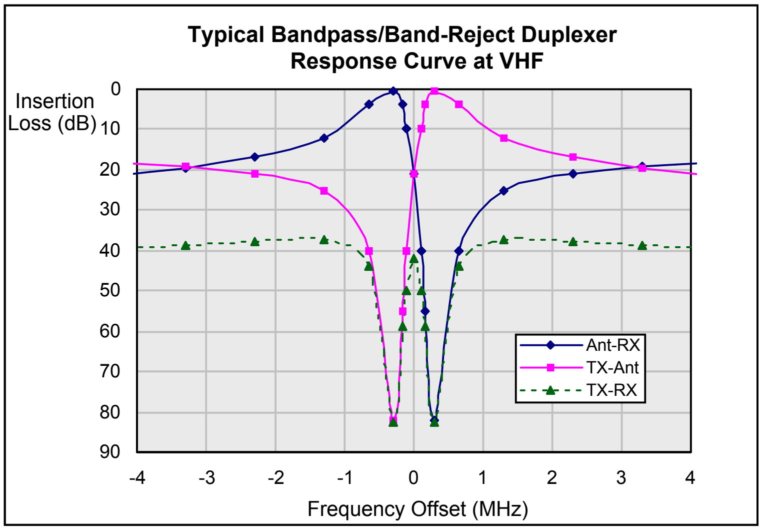

Figure 10. A typical frequency response for a duplexer. "Ant-RX" is the insertion loss between the antenna and receiver. "TX-Ant" is the insertion loss between the transmitter and antenna. "TX-RX" is the isolation between transmitter and receiver.

Figure 10 reveals a limitation of using duplexers (and, hence, a single antenna) at a multichannel site. The useful passband and notchband are relatively narrow, and only in cases where a multichannel site has just a few channels will a duplexer (and single antenna) be a viable alternative.

Figure 11 shows the aggregate transmission bandwidth of a multichannel RF signal utilizing WCTF's band plan, along with the corresponding "guardband" between the most closely-spaced transmit and receive frequencies. For example, an individual site complying with the proposed WCTF band plan* but only using six of the possible eight channels, would require a transmission bandwidth of $(6 - 1) \times 75 + 12.5 = 387.5$ kHz. The separation between the lowest-frequency transmit channel and the highest-frequency receive channel would be $720 - 387.5 = 332.5$ kHz.

* 720-kHz TX-RX frequency split, 75-kHz channel spacing.

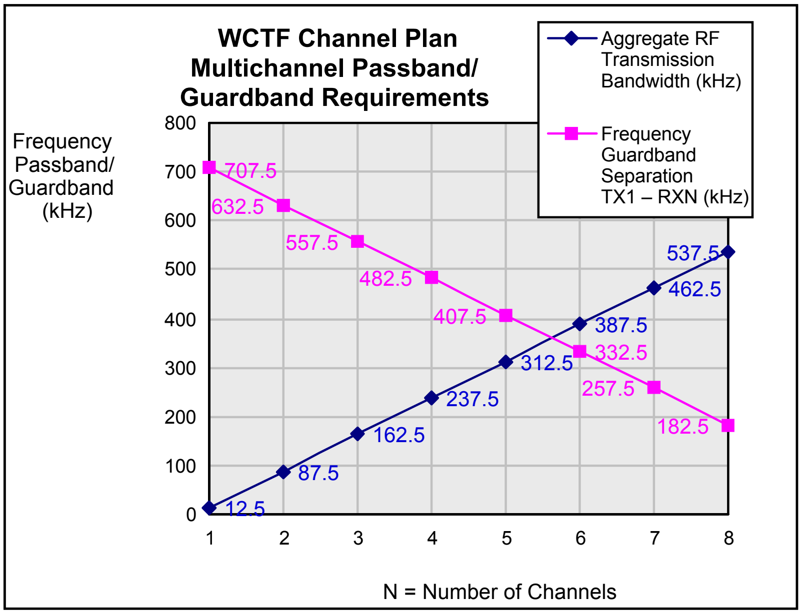

Figure 11. Passband and guardband constraints required of a duplexer versus the number of channels, using the WCTF band plan.

In summary, a duplexer approach, which would permit collocated repeaters and (duplex) base station transceivers to share a single antenna, is suitable only for sites having just a few channels. A greater number of collocated channels requires a different solution.

5.4 Separate Antennas for Transmit and Receive

Separate antennas, one dedicated for transmit and the second dedicated for receive, offer distinct advantages over a single antenna/duplexer approach. Figure 12 shows the amount of isolation between two vertically-polarized, vertically-collinear, dipole antennas versus the vertical separation distance between them. Additional isolation between transmitters and receivers can be obtained by using passband filters and taking advantage of their frequency selectivity.

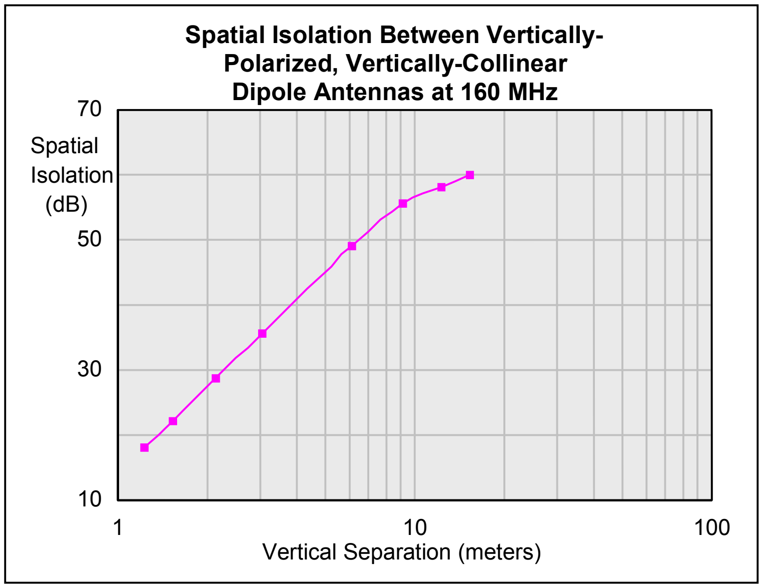

Figure 12. Spatial isolation between two vertically-polarized, vertically-collinear dipole antennas.

Disadvantages to using separate antennas for transmit and receive include:

- antenna tower heights must be much taller to accommodate the second antenna,
- taller towers may be prohibited by local zoning ordinances,
- an additional antenna may be impractical on shared antenna tower facilities because
  - there may be no place on the tower to affix it such that there is adequate isolation between it and other services' antennas,
  - the impact of additional antennas on the antenna tower's wind/ice-loading design margins may be a concern, and
- greater cost.

5.5 Intermodulation Interference

The preceding sections have presented several approaches to allow multiple transmitters and/or receivers to share a common antenna. However, collocated transmitters and receivers might be subject to intermodulation interference.

5.5.1 Determining Intermodulation Frequencies

Intermodulation interference arises from the unintentional mixing of two or more RF signals in a transmitter or receiver. Consider a hypothetical multichannel site with channel assignments belonging to one of the trunk groups* of the proposed WCTF band plan. If channels 2, 3, 4, 7, and 8 are simultaneously transmitting, one possible intermodulation frequency is

$$\begin{aligned} f_{IM} &= f_{E2(tx)} + f_{E3(tx)} + f_{E4(tx)} - f_{E7(tx)} - f_{E8(tx)} \\ &= 161.0400 + 161.1150 + 161.1900 - 161.4150 - 161.4900 \\ &= 160.4400 = 160.4700 - 0.03 = f_{E4(rx)} - 0.03 \end{aligned} \quad (1)$$

In this example, an unwanted intermodulation frequency is 30 kHz away from the assigned frequency of the channel 4 receiver. Performing this calculation using *any other* trunk group gives a similar result. That is, when channels 2, 3, 4, 7, and 8 at any collocated site are simultaneously transmitting, one of the possible intermodulation frequencies will be 30 kHz less than the tuned frequency of that site's channel 4 receiver. That is because the channel spacing within the trunk group and the transmit-receive frequency split of each channel-pair are the same from one trunk group to the next.

This example would seem to imply that since the intermodulation product is more than two receiver bandwidths away (30 ÷ 12.5) from the assigned frequency of the "channel 4" receiver, the "channel 4" receiver would not be subjected to intermodulation interference. However, the finite bandwidths of each channel, due to the modulation of each of the contributing carrier frequencies, have not been considered.

Holbeche [9] has modeled a modulated carrier as a "cluster" of many unmodulated sinusoids. The cluster occupies a finite bandwidth. Computing the intermodulation interference components arising from different combinations of these individual sinusoidal components, Holbeche showed that the bandwidth of the resulting cluster of interference components equals the sum of the bandwidths of each contributing signal. So if several signals each occupy a bandwidth of 12.5 kHz, the bandwidth of a 3rd-order intermodulation product is 3×12.5 kHz. Similarly, the bandwidth of a 5th-order intermodulation product is 5×12.5 kHz. It is therefore conceivable that modulated carriers could cause interference to a receiver while unmodulated carriers would not.

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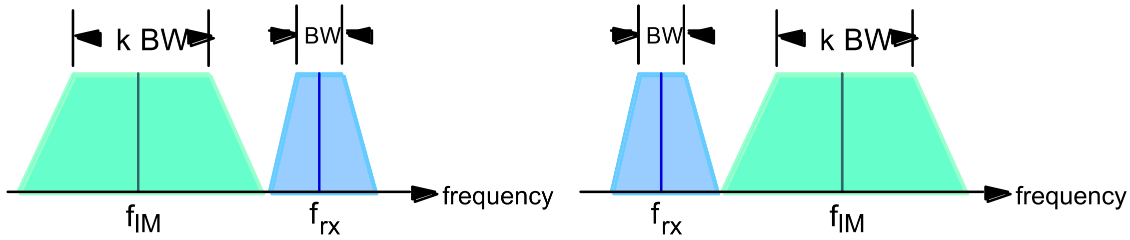
* Recall that within the railroad VHF spectrum, WCTF's proposed band plan has ten interstitially-placed trunk groups. These trunk groups are identified as trunk groups "A" through "J" and the important features of these interstitial trunk groups were depicted in Figure 2. Although any one of these ten trunk groups could be chosen to explain the concepts presented in the subsequent discussion, the frequency assignments corresponding to trunk group "E" frequencies (160.965/160.245, 161.040/160.320, 161.115/160.395, … , 161.490/160.770 MHz) have been arbitrarily selected to explain these concepts.



(a) (b)

Figure 13. Bandwidth of intermodulation products: (a) $f_{rx} > f_{IM}$, (b) $f_{IM} > f_{rx}$. If the spectrum of the intermodulation product overlaps into the passband of a receiver, there is a potential for intermodulation interference.

Figure 13 shows two cases where the frequency of an intermodulation product is nearly the same as a receiver's frequency. A "k^{th}-order" intermodulation product possesses a bandwidth "k" times greater than the fundamental bandwidth and is identified by the term "kBW." To avoid intermodulation interference, the difference between an intermodulation product's center frequency and the receiver's tuned frequency should be greater than the sum of half the intermodulation product's bandwidth and half the receiver's passband, i.e.,

$$\left| f_{IM} - f_{rx} \right| > kB_t/2 + B_r/2 \tag{2}$$

where k is the order of the intermodulation product, B_t is the transmitter bandwidth, and B_r is the receiver bandwidth.

Reconsider, then, Equation (1). But now, assume that "significant" spectral components exist ±5.1 kHz away from each of the assigned transmitter channel frequencies. Following Holbeche's model, a possible intermodulation frequency arising from the spectral components of these modulated carriers could now occur at

$$\begin{aligned} f_{IM} &= (161.04 + .0051) + (161.115 + .0051) + (161.19 + .0051) \\ &\quad - (161.415 - .0051) - (161.49 - .0051) \\ &= 160.4655 = 160.47 - .0045 = f_{E4(rx)} - .0045 \end{aligned}$$

This says that there is a possibility of fifth-order intermodulation interference, arising from the unintentional mixing of channels 2, 3, 4, 7, and 8, occurring only 4.5 kHz away from the assigned frequency of the "channel 4" receiver. This intermodulation interference would fall within a TIA-102 receiver's ±6.25-kHz passband.

The above example gives just one of many possible conditions for generating intermodulation interference. In a more general context, intermodulation frequencies are caused by nonlinearities in transmitter or receiver components. These nonlinearities distort the intended signal $y = A_1 x$, and can be represented mathematically by including additional power series expansion terms, i.e.,

$$y = A_0 + A_1 x + A_2 x^2 + A_3 x^3 + \ldots + A_n x^n \qquad (3)$$

where y represents the output response of a radio to an input signal x and the terms A_n are integer coefficients. When x is a sum of sinusoidal signals, trigonometric expansions of the power terms of x (x^2, x^3, etc.) show that new signals are formed at new frequencies not present in the first-order x term. These new frequencies occur at values given by different combinations of sums and differences of the original fundamental frequencies in x, and their harmonics. So, in general, intermodulation frequencies are given by

$$\begin{aligned} f_{IM} &= \pm k_1 f_1 \pm k_2 f_2 \pm k_3 f_3 \pm \ldots \pm k_N f_N \\ &= \pm k_1 f_1 \pm k_2 (f_1 + \partial f) \pm k_3 (f_1 + 2\partial f) \pm \ldots \pm k_N (f_1 + (N-1)\partial f) \end{aligned} \qquad (4)$$

where each k_j is an integer coefficient designating the "k^{th}" harmonic of frequency f_j, and ∂f is the channel spacing between the collocated radio repeater channels.

For the condition where the sum of the harmonic coefficients $\Sigma(\pm k_j) = 1$, the expression for f_{IM} becomes

$$f_{IM} = f_1 \pm k_2\ \partial f \pm k_3\ 2\partial f \pm \ldots \pm k_N (N-1)\ \partial f \qquad (5)$$

Equation (5) indicates that for equal channel spacings, a fixed transmit/receive frequency split, and a particular set of channel harmonic coefficients [$\boldsymbol{k}$], the frequency of the intermodulation interference signal maintains a fixed arithmetic offset from the base channel frequency f_1. Whether the intermodulation interference actually occurs or not depends on many other factors such as:

- individual transmitter power spectral densities,
- ambient receiver noise level,
- intermodulation rejection specification of the receiver,
- amount of attenuation that transmitter signals experience as they propagate through the various hybrids, cavities, postselectors, and preselectors in the transmitter-to-receiver path, and
- the probability that the offending transmitters are simultaneously transmitting.

5.5.2 Amplitudes of Intermodulation Distortion Products

To estimate the absolute power levels of intermodulation interference, the equations shown in Figure 14 are used. In this figure, the ordinate is "output power minus amplifier gain," and the abscissa is input power. The two heavy, intersecting sloped lines are the first-order (fundamental) "output power minus amplifier gain" vs. input power (1:1 log-log slope) and the third-order intermodulation distortion "output power minus amplifier gain" vs. input power (3:1 log-log slope).

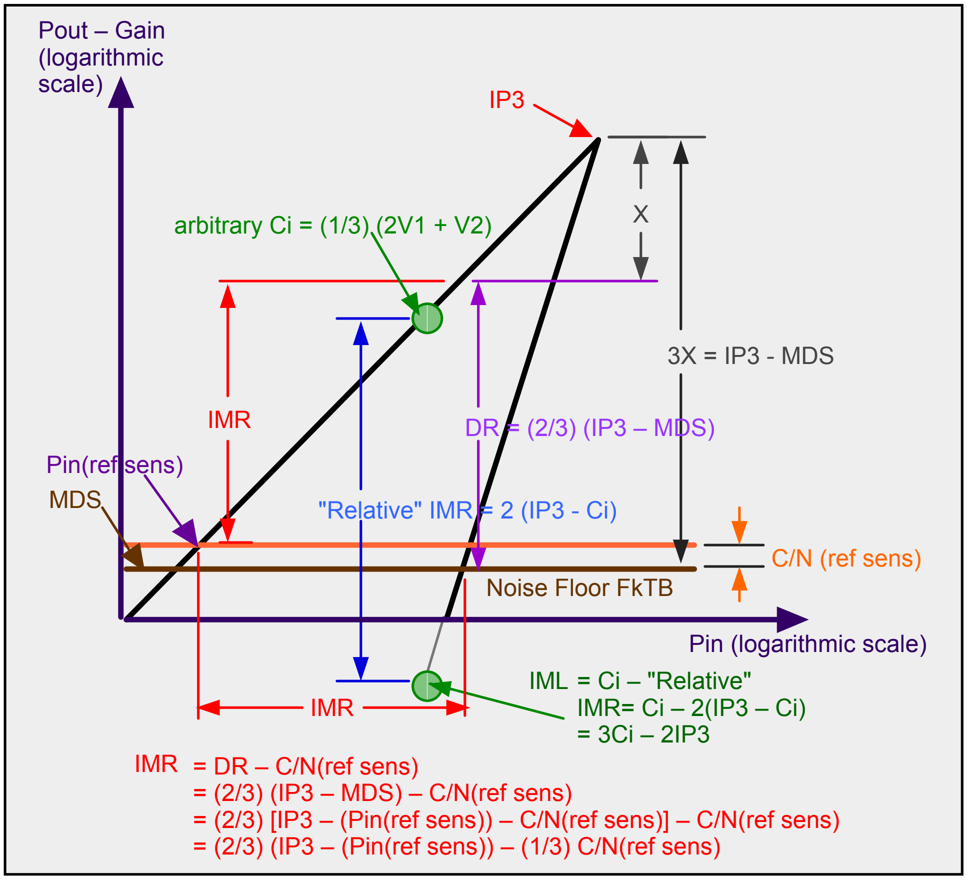


Figure 14. Amplifier intermodulation performance diagram.

The amplitudes of these intermodulation interference signals are determined by expanding the power series in Equation (3) term by term. Doing this, one can show that for every 1-dB increase in the amplitude of the first-order power terms, the third-order distortion terms increase by 3 dB. That is why the first- and third-order responses in Figure 14 have 1:1 and 3:1 log-log slopes, respectively.

In practice, the amplifier's first- and third-order responses will saturate (the two lines will flatten out to horizontal, indicating constant power output with continued increases in input power) before the lines intersect. However, for analysis purposes, the responses are linearly extrapolated to the point where they do intersect. This intersection point is known as the third-order intercept point, and is denoted by *IP3* (or *IIP3* to emphasize third-order *input* intercept point).

$P_{in(ref\,sens)}$ is the input power required to achieve some measure of radio receiver sensitivity, such as 5% BER or 12-dB SINAD. $C/N_{(ref\,sens)}$ is the corresponding input carrier-to-noise ratio when the specified reference sensitivity criteria has been achieved. The intermodulation rejection (IMR) ratio is the ratio between the power level of an off-channel interferer to the power level of the desired channel $P_{in(ref\,sens)}$, when two equal-power off-channel interferers introduce a distortion product whose power is equal to the level of the noise threshold. Other quantities, graphically depicted in Figure 14, include dynamic range *DR*, and minimum discernable signal *MDS*.

When two RF carriers, at power levels C_i that are "*X*" dB less than the receiver's IP3, create intermodulation interference, the third-order intermodulation level (IML) is "*3X*" dB below the receiver's IIP3, at a power level of

$$IML_{(2\text{-}signal)} = 3C_i - 2IIP3 \tag{6}$$

In general, similar plots can be drawn for any "nth-order" intermodulation process, with the nth-order intermodulation response curve possessing an n:1 log-log slope, and with an IML "$n \times X$" dB below the *n*th-order intercept point, at "$nC_i - (n\text{-}1)IIP3$" dBm. Usually, third-order intermodulation products have the greatest IMLs and are the most troublesome. Therefore, the focus of the following analysis is directed to the third-order intermodulation interference problem.

Expanding the third-order term A_3x^3 in Equation (3), when x is the sum of two sinusoids at frequencies f_1 and f_2, gives rise to several terms, one of which is

$$(3/4)\, A_3\, (V_1^2 V_2)\, \cos(2\pi\, (2f_1 - f_2)t) \tag{7}$$

The value of coefficient A_3 can be inferred from an IMR ratio measurement. The standard measurement procedures [6, 7] use two equal-amplitude adjacent-channel signals to create intermodulation interference. When the power of the intermodulation distortion product *IML* just equals the value of the noise, the voltage amplitude V_i of each interfering signal will exceed the reference sensitivity level by an amount equal to the IMR ratio. That is, when

$$IML = (2V_1 + V_2) + 20 \log ((3/4)A_3) = N = (P_{refsens} - C/N_{refsens}) \quad (dBm) \tag{8}$$

then

$$V_1 = V_2 = P_{refsens} + IMR \quad (dBm) \tag{9}$$

When the two interfering signals have unequal amplitudes, the "equivalent" interfering carrier power is

$$C_i = (1/3)(2V_1 + V_2) \quad (dBm) \tag{10}$$

because the resulting intermodulation distortion power *IML*, due to unequal amplitudes V_1 and V_2 on the adjacent and alternate channels, is the same as if V_1 and V_2 were both equal to C_i.

Receiver performance IMR specifications are measured under laboratory conditions using two (adjacent and alternate channel) equal-amplitude signals. However, third-order intermodulation interference can also arise from three signals. When x in Equation (3) is a sum of three sinusoids, the cubic A_3x^3 gives rise to new terms not present in the cubic expansion of a sum of two sinusoids. One of these new terms is

$$(3/2)A_3(V_1V_2V_3)cos(2\pi\,(f_1 + f_2 - f_3)t) \qquad (11)$$

The intermodulation product component described by Equation (11) arises from the "triple beat" of three signals at frequencies f_1, f_2, and f_3. Its leading coefficient term, $(3/2)A_3$, is 6 dB greater than equation (7)'s coefficient $(3/4)A_3$, because

$$20\log((3/2)\,A_3) - 20\log((3/4)\,A_3) = 6\ dB$$

When the amplitudes of these three (adjacent, alternate, and tertiary) signals are the same as the values of the adjacent and alternate channel amplitudes used in the usual "$2V_1 - V_2$" IMR test, the resulting intermodulation distortion term at frequency $(f_1 + f_2 - f_3)$ is 6 dB greater than the intermodulation distortion term at frequency $(2f_1 - f_2)$. Thus, the power level of the intermodulation distortion, due to three signals, is

$$IML_{(3\text{-}signal)} = IML_{(2\text{-}signal)} + 6 = 3\,C_i - 2\,IIP3_{sys} + 6 \qquad (dBm) \qquad (12)$$

When the three interfering signals have unequal amplitudes, the "equivalent" interfering carrier power is

$$C_i = (1/3)(V_1 + V_2 + V_3) \qquad (\,dBm) \qquad (13)$$

because the resulting intermodulation distortion power

$$IML_{(3\text{-}signal)} = IML_{(2\text{-}signal)} + 6 = (P_{ref\,sens} - C/N_{ref\,sens}) + 6 \qquad (dBm) \qquad (14)$$

due to unequal amplitudes V_1, V_2, and V_3, is the same as if V_1, V_2, and V_3 were each set equal to C_i.

As an example, assume an adjacent channel power of -32 dBm, an alternate channel power of -37 dBm, and a tertiary power level of -48 dBm, incident upon a receiver whose third-order input intercept point is +4 dBm.

The equivalent interfering carrier power

$$C_i = (1/3)\,((-32) + (-37) + (-48)) = -39\text{ dBm}$$

The resulting interference at one of the "triple-beat" frequencies is

$$IML_{(3\text{-}signal)} = 3 \times (-39) - 2 \times 4 + 6 = -119\text{ dBm}$$

This level of intermodulation interference is roughly the same as the value of the ambient noise level that might be expected at the receiver output.

The procedure culminating in Equations (3) – (14) could be extended to account for quartic- and higher-order products, A_4x^4, A_5x^5, etc., although the algebraic and trigonometric manipulations become unwieldy in short order.

5.5.3 Composite Receiver System Noise Figure and Intercept Point

The above discussion has investigated intermodulation interference created within a receiver. It did not consider the effects of preceding a receiver with a multicoupler. Collocated repeaters sharing a common receive antenna will have an amplifier preceding the RF splitter to compensate for the losses introduced by splitting the multichannel signal into several distinct receiver paths. A multicoupler will degrade the composite IIP3 and noise figure of the system, and if the preamplifier gain is set too high, the multicoupler will amplify the interfering carriers to such a high level that intermodulation interference is unavoidable. If its gain is set too low, the composite system noise figure will be larger than desired, degrading the system's overall sensitivity.

System noise figure F_{sys} is given by

$$F_{sys} = 10 \log (f_1 + (f_2 - 1)/g_1 + (f_3 - 1)/g_1g_2 + \ldots) \qquad (15)$$

where F_{sys}, the composite system noise figure, is expressed in decibels, and f_i and g_i correspond to the noise factor* and gain of the individual stages comprising the cascaded chain, which are expressed linearly rather than in decibels.

Sagers [10] derived the third-order input intercept point of a cascaded network as

$$1/(IIP3)_{sys} = 1/(IIP3)_1 + g_1/(IIP3)_2 + g_1g_2/(IIP3)_3 + \ldots \qquad (16)$$

where the intercept points $IIP3_i$ and gains g_i are expressed linearly (in milliwatts and as a dimensionless multiplicative factor, respectively).

Consider the following example: Each receiver at a multichannel site has a noise figure of 9 dB and third-order intercept point of +4 dBm. The receivers share a common antenna, so a multicoupler precedes them. The multicoupler's distribution preamplifier has a noise figure of 4 dB, a gain of 25 dB, and a third-order input intercept point of 9 dBm. The multicoupler's RF splitter has an insertion loss† of 17 dB. From Equation (15), the composite system noise figure is

$$F_{sys} = 10 \log \left(10^{\frac{4}{10}} + \frac{10^{\frac{17}{10}} - 1}{10^{\frac{25}{10}}} + \frac{10^{\frac{9}{10}} - 1}{10^{\frac{25}{10}} \cdot 10^{-\frac{17}{10}}}\right) = 5.8 \text{ dB}$$

---

* Noise figure F is related to noise factor f by $F = 10 \log f$.

† The noise figure of a passive, linear, and isotropic device is equal to its amount of attenuation.

This is a 3.2-dB improvement over the noise figure of a single receiver. But from Equation (16), the composite third-order intercept point is found to be

$$IIP3_{sys} = -10log\,(\frac{1}{10^{\frac{9}{10}}} + \frac{10^{\frac{25}{10}}}{10^{+\infty}} + \frac{10^{\frac{25}{10}} \cdot 10^{-\frac{17}{10}}}{10^{\frac{4}{10}}}) = -4.2\ \ dBm^{*}$$

This is 8.2 dB *worse* than the intercept point of the receiver alone. From Equation (12), the level of the intermodulation interference at the final receiver is therefore

$$IML = 3 \times (-39) - 2 \times (-4.2) + 6 = -102.6\ \ dBm$$

The net effect is that the multicoupler has amplified the interference to unacceptably high levels, from -119 dBm to -102.6 dBm. Either the multicoupler's gain must be reduced, or the levels of interference present at the receiver site must somehow be suppressed.

5.5.4 Conclusion

Will intermodulation interference be a problem with multichannel sites assigned frequencies according to the proposed WCTF band plan? If the local environment is such that high-power interference signals are likely to be present, intermodulation interference will be unavoidable unless the system's sensitivity is significantly degraded. This, of course, means that weak signals may not be detected. On the other hand, if lower transmitter power levels can still provide adequate geographical coverage, or if local ambient noise levels and received interference signal strengths are not excessive, then a system design can utilize the proposed band plan. Candidate sites will have to be evaluated on a case-by-case basis.

6 SIGNAL COVERAGE ANALYSIS

The radio coverage area, for a given level of speech intelligibility and service reliability, depends on the effective isotropic radiated power (EIRP) of the transmitter, signal fading and shadowing margins, receive antenna gain, and receiver sensitivity. EIRP is equal to the transmitter output power plus the transmitter antenna gain minus the miscellaneous transmitter system losses (losses through combiners, duplexers, coaxial cable, etc.). A typical mobile antenna will exhibit a gain on the order of 3 dBi, and a typical two-element dipole array will exhibit a gain of 7 dBi. Typical receiver sensitivities for various levels of speech intelligibility were presented in Section 4.1. The remaining parameters, signal fading and shadowing, are considered in the following discussion.

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* The third-order intercept point of a passive, linear, and reciprocal device is, by definition, infinite, since such a device can generate no intermodulation products. Therefore the denominator of the second term inside the parentheses is $10^{+\infty}$.

6.1 Land Mobile Communications Propagation Models

The Hata-Okumura propagation model is commonly used by the LMR community to predict radio coverage [11]. This model is based upon the measurements and empirical prediction curves of Okumura, et al. [12], and is easily implemented using a personal computer spreadsheet program. Hata's model is limited to path distances of 20 km, although Okumura's prediction curves were provided out to 100 km. Others have extended the model to paths up to about 100 km, as in the Hata-Davidson model [4]. The Hata model and Okumura curves considered several different operating environments: (1) urban, (2) medium/small city, (3) suburban, (4) quasi-open, and (5) open. The model assumes that the base station antenna is well above the heights of surrounding buildings for the given environments. Stueber [13] observed that U.S. suburban environments are more like the quasi-open environments defined by Okumura. For railroad operations in cities, the suburban and quasi-open environments are probably the best choices of the Hata model selections.

Other more sophisticated modeling tools are available to communications systems designers. ITS' Communications System Performance Model* (CSPM) incorporates the Longley-Rice Irregular Terrain Model (ITM) algorithm and utilizes digital terrain elevation data to define the terrain profile along the signal propagation path. CSPM produces maps showing signal strength contours within a user-defined geographic area. Alternatively, data can be output in a tabular textual format. Many antenna system manufacturers and engineering consultants have similar computer programs to design their customers' LMR systems, and commercial implementations of the ITM algorithm can be obtained from various sources.

6.1.1 Median Propagation Loss from the Hata-Davidson Model

Figure 15 shows the predicted median propagation loss within the five environments described above for the situation where the base station antenna height is 33 meters, the mobile unit antenna height is 5 meters, and the frequency is 161 MHz. For this given situation of frequency and antenna heights, the Hata-Davidson Model predicts virtually identical attenuations for the urban and medium/small city environments; the curves overlay very closely and cannot be readily resolved.

Two additional curves are also shown on the plot, where the path loss with distance is $10\, n \log_{10}(distance)$ and n is the path loss exponent. The first additional curve is for free-space loss where the path loss term $n = 2$ and the second additional curve is for a path loss term where $n = 3.6$.

A third additional curve is also shown on the plot. This curve shows the loss predicted by ITM over a particular path in the central United States. At distances exceeding about

* More detailed information regarding internet access to the CSPM program can be obtained from the Institute for Telecommunication Sciences, telephone (303) 497-5301.

20 km, variations and undulations in the predicted propagation loss are seen. This behavior is not unexpected since the terrain profile is irregular.

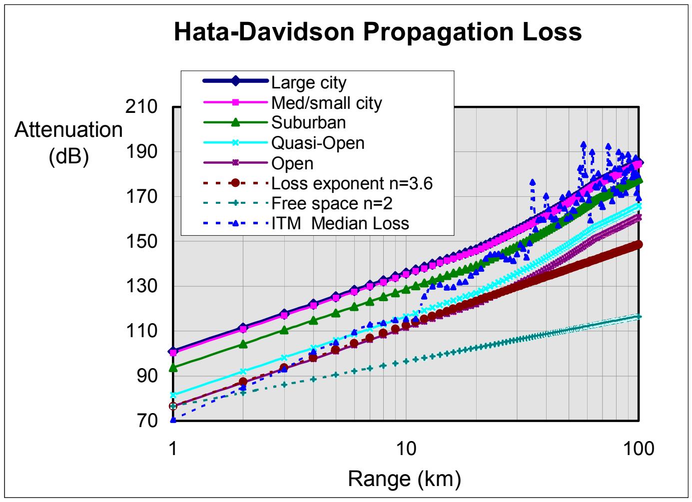

Figure 15. Hata-Davidson, exponential power loss, and ITM models of propagation loss for mobile communications at 161 MHz. Base station antenna height is 33 meters and mobile unit antenna height is 5 meters.

For the purposes of this discussion, the important feature of the ITM prediction is that its propagation path loss behavior obeys the same general trend as the more generic Hata-Davidson model. In addition, the Hata-Davidson curves have a common slope that indicates a path loss exponent of approximately 3.6. Lee [14] has provided examples of measurements where *n* varies over 3.05 (an urban area in Japan), 3.68 (an urban area in Philadelphia), 4.31 (an urban area in Newark), 3.84 (in suburban areas), and 4.35 (in an open area). Thus, the Hata model should be a reasonable estimator with its path loss exponent set to about 3.6.

6.2 Variability about the Median Propagation Loss

The Hata-Davidson model provides a prediction of the median loss versus distance for a given frequency and base station and mobile unit antenna heights. Due to blockage of the signal caused by buildings or terrain plus the reflection of the signal by other buildings and terrain features, the signal as measured at a particular distance is not constant [15]. Multipath conditions experienced by mobile units in motion can also cause Rayleigh fading of the radio signal [15].

6.2.1 Lognormal Shadowing

If the received signal strength were measured at discrete locations along a constant radius surrounding a particular transmitting site, the signal level would be found to vary in a lognormal manner. Figure 16 shows the probability that the actual received signal would vary from the median level by a given amount, for three different values of standard deviation. From the statistics of a lognormal distribution, 68% of all the actual received signals will be no more than one standard deviation from the median signal level.

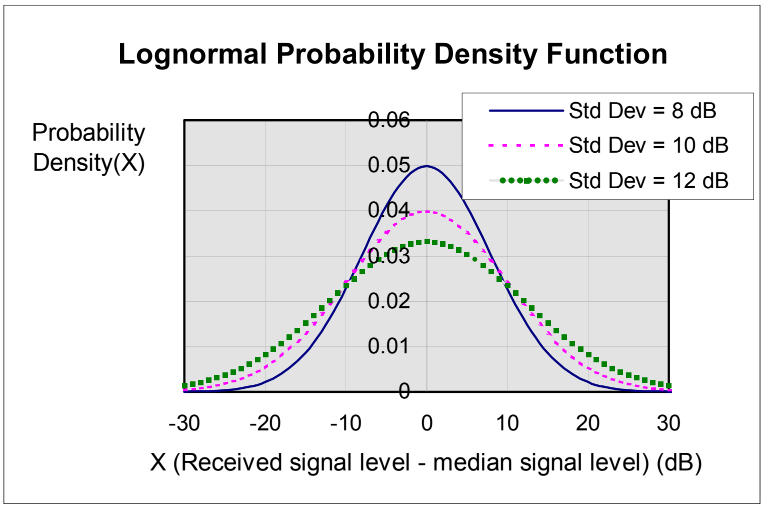

Figure 16. Probability distribution of the difference between the actual received signal level and the median signal level, for three standard deviations of the lognormal signal.

Figure 17 shows the cumulative probability of the actual received signal minus the median signal level for three different standard deviations. The system design needs to be based upon the standard deviation of the signal variability where the system is to be used. Typical values of standard deviation for mobile communications at VHF are about 5 - 12 dB in a variety of urban, suburban, and open environments [14]. A reasonable value, then, for lognormal standard deviation is probably about 8 dB or so for railroad environments.

Assuming an 8-dB standard deviation in signal strength, suppose that the system design specification requires an adequate communications link be available for at least 90% of the times and at 90% of the locations that the mobile user depresses the push-to-talk switch. For a lognormal distribution, the required margin above the median signal level can be found from

$$Prob(X \geq K\sigma) = 1 - {}^{1}/_{2}\, erfc\,({}^{K}/\sqrt{2}) \qquad (17)$$

where *erfc(·)* is the complementary error function and is given by

$$\tfrac{1}{2}\, erfc(\upsilon\,) = \int_{\upsilon}^{+\infty} e^{-\xi^2}\, d\xi \qquad (18)$$

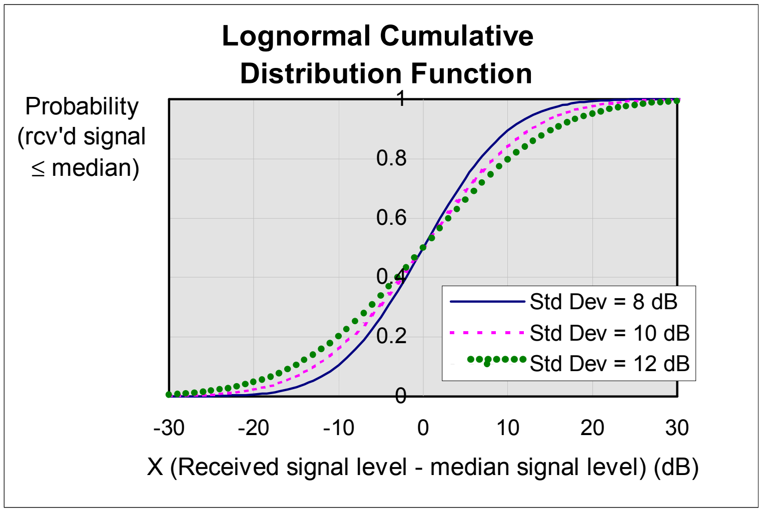

Figure 17. Cumulative probability that the actual received signal level will differ from the median signal level by at least a specific amount, for three standard deviations of the lognormal signal.

The *erfc(·)* arises from integrating the lognormal probability density function of a Gaussian-distributed random variable ξ. X is the difference between the received signal level and the median signal level and σ is the standard deviation of X. For the required $Prob(X \geq K\sigma) = 90\%$, K is found to be 1.28 so X, the margin, would need to be about 10.25 dB (1.28 × 8 dB). Suppose instead the system reliability was specified to be 95%, then K would be 1.65 and X would need to be 13.2 dB (1.65 × 8 dB).

To see how this margin is used in the design of the system, assume the Hata-Davidson model predicts a propagation loss of 150 dB at a distance of 24 km. If the *available excess system link margin* equals the median propagation loss, then an adequate signal level will exist 50% of the time at 50% of the locations along a circular contour 24 km away from the transmitter. By subtracting 10.25 dB from the available excess system link margin, the design will assure that 90% of these locations will have an adequate signal level. Subtracting 13.2 dB from the available excess system link margin would assure that 95% of these locations will have an adequate signal level.

6.2.2 Rayleigh Fading

Rayleigh fading describes the phenomenon of signal fluctuations caused by multipath effects while a mobile unit is in motion. The probability density function of a signal whose envelope (voltage) r obeys Rayleigh fading statistics is

$$p_r(r) = \frac{r}{\gamma} e^{-\frac{r^2}{2\gamma}} \qquad (19)$$

where γ is the mean power of the signal. Equation (19) is graphed in Figure 18.

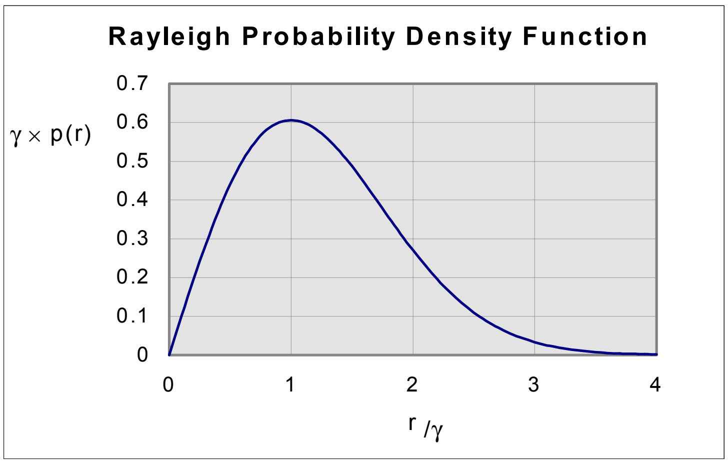

Figure 18. Normalized probability distribution of the magnitude of a Rayleigh-faded signal.

The probability density function of this same Rayleigh-faded signal can be expressed in terms of the signal power s by equating incremental areas under the probability density curves $p_s(s)ds$ to $p_r(r)dr$, i.e., $p_s(s) = \frac{p_r(r)}{\left|\frac{ds}{dr}\right|}$, yielding the exponential probability density function

$$p_s(s) = \frac{1}{2\gamma} e^{-\frac{s}{2\gamma}} \qquad (20)$$

This function is graphed in Figure 19.

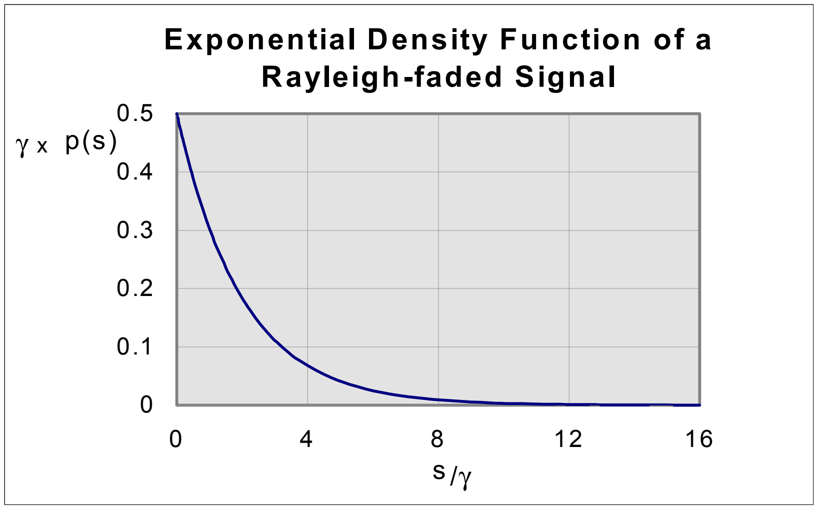

Figure 19. Normalized probability distribution of the power of a Rayleigh-faded signal.

Hess [15] and Parsons [16] further developed Equation (20) to express the probability density function in terms of decibels* rather than watts. The result is

$$p_S(S) = \frac{1}{2\kappa\,\gamma} e^{\left(\frac{S}{\kappa} - \frac{1}{2\gamma} e^{\frac{S}{\kappa}}\right)} \tag{21}$$

where $S = 10\ log(s) = \dfrac{10}{ln(10)}\ ln(s) = \kappa\ ln(s)$, and the mean signal power γ is expressed linearly as before. The authors [15,16] also showed that the mean of the received power in a Rayleigh-fading environment is about 2.51 dB less than for the static case, at a value of $\kappa\ ln(\gamma) - \kappa C$ dB,† with a standard deviation σ_r equal to $\dfrac{\kappa\,\pi}{\sqrt{6}}$ dB, or about 5.57 dB, with σ_r being independent of the value of mean power.

6.2.3 Suzuki Distribution

The superposition of Rayleigh fading onto lognormal shadowing yields a probability density function known as the *Suzuki Distribution* [16]. When determining the statistics of a Suzuki-distributed signal, Equations (20) and (21) must be written in a slightly different form. In the land-mobile radio environment, the mean power γ‡ of a Rayleigh-faded signal is itself a random variable, exhibiting the lognormal-shadowing behavior described

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* $S = 10\ \log s$

† Euler's constant $C \approx 0.577215665$

‡ or $\Gamma = 10\ \log \gamma$ (dB)

previously. Therefore, Equations (20) and (21) are rewritten as *conditional* probabilities $p_{s|\gamma}(s|\gamma)$ and $p_{S|\Gamma}(S|\Gamma)$. The composite probability density function $f_s(s)$ becomes $\int_{-\infty}^{+\infty} p_{s|\gamma}(s|\gamma) p_s(s) d\gamma$, where the second term describes the lognormal behavior and the integration corresponds to averaging the first (Rayleigh) term weighted by the lognormal distribution. Hess [15] gives the Suzuki probability density function of power as

$$f_s(s) = \frac{1}{\sqrt{2\pi\sigma^2}} \int_{-\infty}^{+\infty} e^{\left(\frac{\gamma}{\kappa} - se^{-\frac{\gamma}{\kappa}} - \frac{(\gamma - \bar{\gamma})^2}{2\sigma^2}\right)} d\gamma \qquad (22)$$

where $\bar{\gamma}$ is the mean value of the non-Rayleigh-faded received power γ, σ is the standard deviation of the lognormal shadowing distribution, and κ is the proportionality constant defined previously.

This probability density function and its resulting cumulative distribution function $F_S(S \leq S_{r(dB)}) = \int_{-\infty}^{S_r} f_s(S)dS$ present significant computational difficulties. Parsons [16] reports that other theoreticians have suggested that subject to certain limitations,* the Suzuki Distribution can be approximated by a lognormal distribution with a mean power level $\bar{S}_{Suzuki} = (\bar{S}_{lognormal} - \kappa C)$ and standard deviation $\sigma_{Suzuki} = \sqrt{\sigma^2_{log\,normal} + \frac{\kappa^2\pi^2}{6}}$, both given in decibels.

6.3 Excess System Link Margin

Based upon the above discussions, the excess system link margins and maximum signal coverage distances can be determined, for both digital and analog modulation formats of a TIA-102 radio system. The excess system link margin indicates how much path loss a radio channel can withstand while providing some specified minimum reliability of service and delivered audio quality out to some distance. The maximum available excess system link margin is given by

$$Available\ margin = EIRP_{TX} + G_{RX} - L_{system} - P_{ref\,sens} - \kappa C - K\sigma_{Suzuki} \qquad (23)$$

Table 1 shows two sets of example calculations, using Equation (23), to determine the available excess system link margins for analog and digital modulation formats. The available excess system link margin is a function of several transmitter and radio channel parameters.† For this example, a DAQ of 3.4 is assumed, which, according to TIA [4],

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* For example, $\sigma_{lognormal} \geq 6$ dB and reliability percentiles bounded within 5% and 95% values.

† These parameters are: (1) the transmitter's effective isotropic radiated power (*EIRP*), (2) the mobile receiver's antenna gain (G_{RX}), (3) miscellaneous losses within the local RF infrastructure (L_{system}), (4) the receiver noise level (FkTB), (5) the carrier-to-noise ratio (*CNR*) required for a specified grade of service, and (6) the contour reliability (which determines $K\sigma$). The constant term κC, discussed earlier, accounts for the reduction in the mean power of the received signal due to Rayleigh fading.

requires a CNR of 17.7 dB. Similar calculations can be done for base-to-portable and mobile/portable-to-repeater scenarios.*

Notice that the TIA-102 digital modulation gives about an 7 dB performance advantage over analog FM. Thus, given the same conditions for both options, the digital modulation should have superior performance to that expected of the analog modulation option. From another viewpoint, for a given level of speech intelligibility, digital modulation should provide coverage out to a further range than analog modulation.

Table 1. Excess System Link Margin Calculations for Analog and Digital Modulation Formats — Base-to-Mobile

Available Excess System Margin	IMBE vocoder transceiver	Analog voice transceiver
Transmitter EIRP = Pt - L_{system} + $G_{antenna}$ (dBm)	46.5	
Receiver antenna gain (dBi)	3.0	
Rx cable losses (dB)	1.0	
Rx equiv noise bandwidth (kHz) (TIA TSB-88 Jan98 p.90)	5.7	10.1
Noise level = kTB (dBm)	-136.4	-133.9
Rx DAQ2/5%BER/12dBS reference sensitivity (dBm)	-113.0	-113.0
CNR for DAQ 2.0 (dB) (TIA TSB-88 Jan 98 p.93)	7.6	5.0
Receiver noise figure (dB)	15.8	15.9
CNR for DAQ 3.4 (dB) (TIA TSB-88 Jan 98 p.93)	17.7	22.0
Rayleigh fading mean value correction κC (dB)	2.5	
Rayleigh fading standard deviation $\kappa\pi/\sqrt{6}$(dB)	5.6	
Lognormal shadowing standard deviation (dB)	8.0	
Contour reliability $P(X \geq K\sigma)=1-(1/2)\mathrm{erfc}(K/\sqrt{2})$ (%)	90.0	
Suzuki fading margin $K\sigma$ (dB)	12.5	
Available Excess System margin (dB)	136.4	129.5

Figure 20 shows the excess system link margin versus distance of the base-to-mobile channel for the two modulation options in two different environments, by subtracting Figure 15's median path loss from Table 1's *available* excess system link margin. The points where each line crosses the abscissa (i.e., where the excess margin equals zero) define the maximum estimated ranges that signal coverage can be expected to provide the specified quality and reliability. The figure shows that the digital system should provide signal coverage out to about 17 km in suburban environments, and out to about 31 km in quasi-open environments. This is approximately 1.4 times the coverage distance that an analog system can provide, for the same level of speech intelligibility.

* For mobile/portable-to-repeater, substitute the value of the mobile/portable transmitter power for the transmitter EIRP, and for portables (inbound or outbound), substitute the value of the portable's antenna gain (nominally about –10 dBi at VHF).

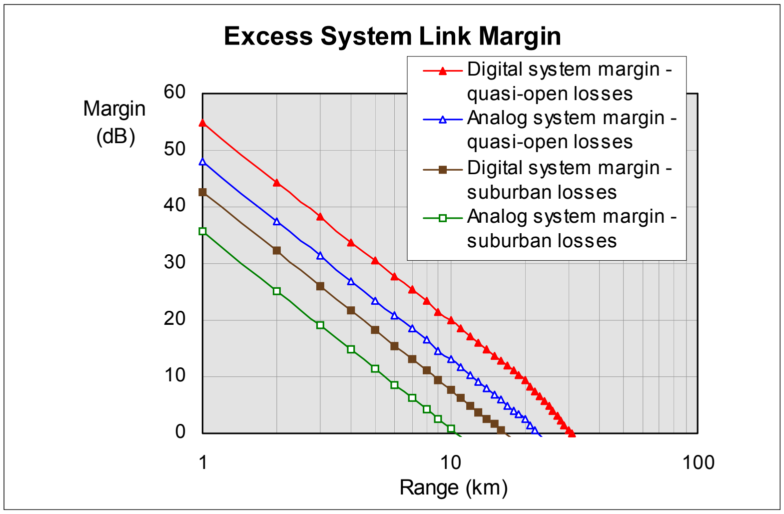

Figure 20. Typical excess system link margin for digital and analog modulation formats in two different terrain environments.

6.4 Cellular Topologies

In regions where many rail lines converge and rail yard activity increases the amount of traffic, cellular topologies can be used to provide the required signal coverage. Typically, a hexagonal shape is used to describe the most basic coverage area. From this shape, several patterns are defined that provide different separations between base station transmitters using the same frequency. The common patterns are found by moving i cells in one direction away from the current cell, turning 60 degrees, then moving j cells to arrive at a new cell. This new cell uses the same frequency as the starting cell. The formula for the frequency reuse factor N, which is equal to the number of cells per cluster, is related to the allowed moves i and j by

$$N = i^2 + ij + j^2 \tag{24}$$

Figure 21 shows several examples of cellular clusters for several different frequency reuse values N.

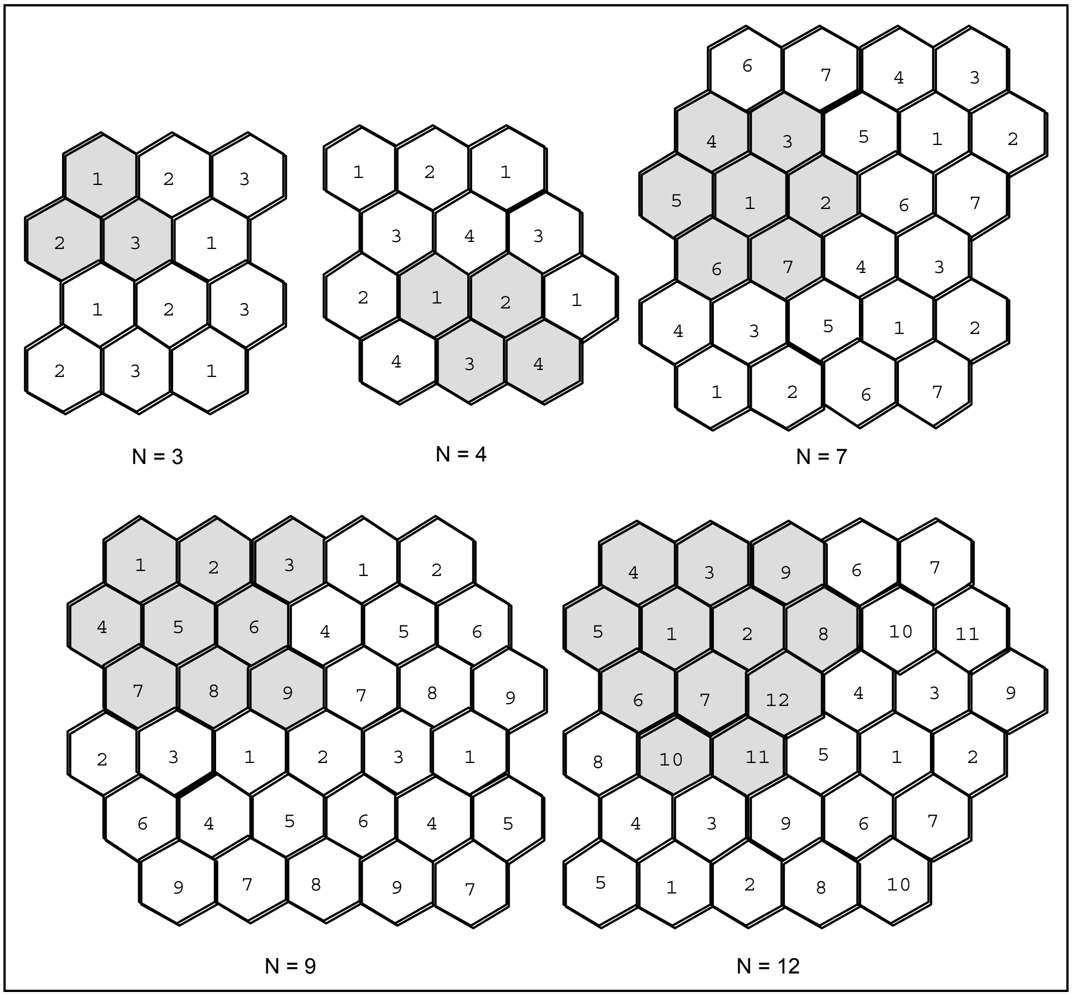

Figure 21. Example cellular cluster designs for several different values of frequency reuse factors N.

The quantity Q is the ratio of the distance D between the centers of cochannel cells and the radius R of a hexagonal cell. It is related to the frequency reuse factor N by

$$Q = D/R = \sqrt{3N} \tag{25}$$

Generalizing Rappaport's approximation for the cochannel carrier-to-interference ratio [17] to account for any path loss exponent n gives

$$\frac{C}{I}_{cochannel} = -10\log\left\{ \frac{2(Q+1)^n + (Q-1)^n}{(Q^2-1)^n} + \frac{(Q+\frac{1}{2})^n + (Q-\frac{1}{2})^n}{(Q^2-\frac{1}{4})^n} + \frac{1}{Q^n} \right\} \tag{26}$$

Equation (26) describes the situation where a mobile unit is at the boundary of its cell and the base stations all have equal characteristics. Table 2 gives the expected $C/I_{cochannel}$, computed using Equation (26), for several different values of path loss coefficient n for the most common cellular patterns such as those depicted in Figure 21.

Table 2. Cellular Cluster Size N and Cochannel Carrier-to-Interference Ratio

i	j	$N = i^2 + ij + j^2$	$Q = D/R = \sqrt{3N}$	C/I (dB) path loss coefficient n			
				2	3	3.6	4
1	0	1	1.73	-7.04	-7.76	-8.35	-8.79
1	1	3	3.00	0.38	4.09	6.22	7.61
2	0	4	3.46	1.94	6.51	9.18	10.93
2	1	7	4.58	4.76	10.86	14.48	16.87
3	0	9	5.20	5.97	12.72	16.73	19.39
2	2	12	6.00	7.33	14.79	19.24	22.19
3	1	13	6.24	7.70	15.36	19.92	22.95
3	2	19	7.55	9.45	18.01	23.12	26.53
3	3	27	9.00	11.05	20.42	26.03	29.77

Cellular patterns incorporating variable-sized cells may be warranted. Reducing cell size (thus increasing the total number of cells encompassing a geographic area) is one way to increase the traffic capacity of radio systems.* Figure 22 shows a possible frequency group reuse plan using variable-size cells. The smaller sized cells would be used where the greatest need for channel capacity exists, i.e., rail centers. Larger-sized cells would be used in areas with reduced channel capacity.

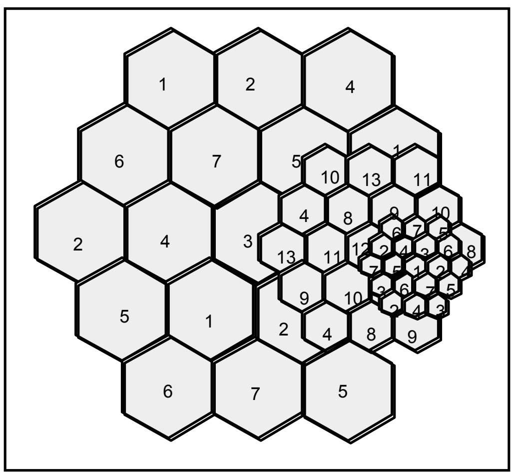

Figure 22. Example variable cell cluster size for a frequency reuse factor N = 7, using 13 frequency groups to control interference in "cell-size transition regions."

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* The ability to dynamically control transmit power level is desirable in order to accommodate smaller cell sizes. Too much transmitter power in small-sized cell configurations causes interference to radios in cochannel or neighboring adjacent-channel cells. Currently, P25 radios do not have the capability to automatically control the transmit power of mobile and portable units from the repeater or base station. However, some measure of protection can be had by controlling antenna heights, beam downtilt, gain, and radiation pattern at the repeater/base station. ITS has suggested automatic power control be incorporated into the Project 25 standards (private verbal communications, TIA-102 standards committee).

6.5 Adjacent-Channel Rejection, Cochannel Rejection, and Frequency Reuse Ratio (Cell Cluster Size)

Typical measured adjacent-channel and cochannel rejection performance of analog-modulated and digitally-modulated radios was presented in Section 4.2. In that section, typical delivered audio qualities of demodulated speech signals were given for a variety of adjacent-channel and cochannel environments. The following subsections discuss how cellular topology and narrow channel spacings affect the intelligibility of the received speech signal.

6.5.1 Adjacent-Channel Rejection

The worst-case situation for determining the effects of adjacent-channel interference on DAQ occurs in the portable-to-base direction, when a lower-powered portable transmitter must compete against higher-powered interference sources. Refer to Figure 23. Two adjacent-channel mobiles, located in neighboring cells at the cell boundaries, are depicted interfering with the reception of a desired signal from a lower-powered portable located within the cell of interest.

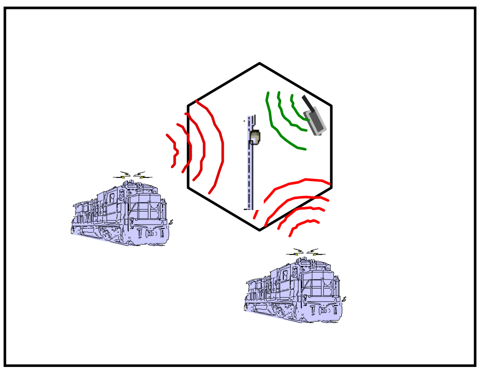

Figure 23. Portable-to-repeater situation: Two adjacent-channel mobiles near the boundaries of neighboring cells interfere with the reception of a desired signal from a lower-powered portable located within the target cell.

When ITS measured the DAQ performance of a TIA-102 radio in the presence of 7.5-kHz offset adjacent-channel interference (Section 4.2), the desired transmitter's power level was first adjusted to invoke a 0.25% BER at the receiver. This provided a signal whose DAQ was superior to 3.4. Interference signals were then introduced to degrade the desired signal's DAQ. When the adjacent-channel interference power was about 9 dB greater than the on-channel signal, the DAQ of the desired speech signal was degraded to a score of 3.4 (Figure 5). This implies that two independent interference signals, one on the lower adjacent channel and the other on the upper adjacent channel, could each be about 6 dB stronger than the intended signal at the receiver, since $10 \log(10^{\frac{6}{10}} + 10^{\frac{6}{10}}) = 9$ dB.

An analysis* such as that performed in Section 6.3 predicts a maximum distance (where the excess system link margin is 0 dB) between portable and base station of about 2 km in suburban environments for a 0.25% BER. Performing a similar analysis for interfering mobiles† to find where the excess system link margin is 6 dB,‡ shows that the interfering mobiles should be no closer than 37 km to the receiving base station. Since, for the assumed conditions of this example, the cell radius is 17 km (Section 6.3), two interfering adjacent-channel mobiles in neighboring cells, offset ±7.5 kHz in frequency from the on-channel portable, might degrade the desired signal's DAQ to something less than DAQ 3.4. However, the railroads may decide that a DAQ less than 3.4 still gives acceptable speech quality, or that this situation occurs infrequently enough that it is not an issue.

6.5.2 Cochannel Rejection

Table 3 gives the cochannel carrier-to-interference (C/I) ratios measured by ITS and relates these values to the corresponding minimum cell cluster sizes *N* from Table 2, for the interference modes and channel spacings specified in Section 4.2.

Table 3. Representative Values of *C/I* and *N* for Different Radio Interference Environments

Frequency Reuse Ratio (Cell Cluster Size) *N* for Path Loss Exponent *n* = 3.6 and Delivered Audio Quality DAQ = 3.4		
Interference Mode	Cochannel	
	C / I (dB)	N
unwanted digital signal interfering with desired digital signal	9.2	7
unwanted analog signal interfering with desired digital signal	18	12
unwanted analog signal interfering with desired analog signal	25	19
unwanted digital signal interfering with desired analog signal	not measured	

The table shows that the number of cells comprising a cluster§ can be less for an all-digital environment than for all-analog or mixed-mode environments. This means that

* Portable: 0.5-watt transmitter EIRP, 1.0-meter antenna height. Base station: 33-meter antenna height, 5-dBd antenna gain. Estimated CNR for 0.25% BER: 31 dB.

† 45-watt transmitter EIRP, 5.0-meter antenna height. Contour coverage reliability: 50%.

‡ From the previous paragraph, two simultaneously occurring adjacent-channel interfering signals can each be 6 dB stronger than the desired signal.

§ Each shaded group of cells in each of the cellular topologies depicted in Figure 21 is a cluster.

more channel-pairs can be assigned to each cell. If "C" channel-pairs are available for use throughout a cluster of cells, then "C/N" channel-pairs can be assigned to each cell. For eighty channel-pairs and $N = 7$, eleven to twelve channel-pairs can be assigned to each cell.

7 TRUNKING AND TRAFFIC LOAD CAPACITY

Collocated repeaters can be configured to significantly increase the amount of supportable communications traffic. Utilizing a concept known as trunking, many more users can be served by sharing a finite number of radio channels, similar to how a finite number of telephone trunk lines connecting telephone central office switches can be shared by a large number of individual telephone customers. This is possible because of the statistics related to any particular user needing access to a telephone circuit, or, in this case, a radio channel. These statistics have been well documented and are known as Erlang traffic queueing theory. A family of curves has been developed that illustrates the gains possible in system capacity by incorporating trunking strategies, and is presented in Figure 24.

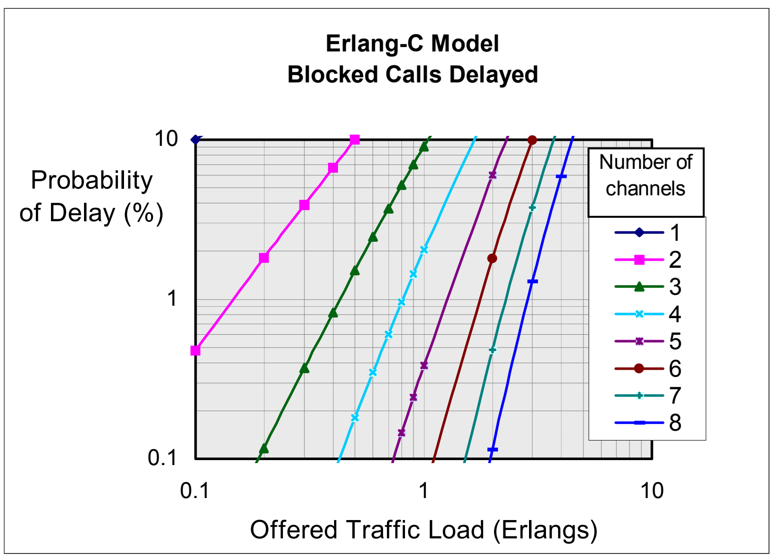

Figure 24. Erlang-C traffic queueing theory relating the traffic load offered to the network to the number of trunking channels.

The unit of measure of traffic intensity is known as the Erlang. The Erlang-C (blocked calls delayed) model is presented here. The model assumes that channel requests arrive randomly and that there is an infinite population of users. A blocked user will be queued up to some predetermined time pending a channel grant. If the user is not assigned to a traffic channel within this period of time, the request is dropped, the user gets a "system busy" indication, and the user is returned to the population pool. If returned to the population pool, the model further assumes that the user is no more likely to initiate a follow-on request for an available channel than any other user in the population pool is to generate a channel request.

To use Figure 24, a maximum acceptable probability of being denied immediate assignment to a traffic channel must be known. For example, assume that a 2% probability of a call not being immediately granted a traffic channel is acceptable. Then, on average, 2% of the time that a user initiates a channel request by depressing the radio's PTT switch, the system would queue the call request because all trunk channels were busy. Find the 2% blocking point on the "Probability of Delay" axis of the graph and move horizontally across to where it intersects the appropriate "No. of Channels" curve. Then move vertically down from this point to the "Offered Traffic Load" axis.

Consider a trunked system with eight channels at each site. One of these channels will be reserved exclusively for control channel functions and the remaining seven channels will be available for message traffic. If a 2% probability of blocking is desired, the total *offered load* that this trunking system could support would be about 2.6 E.

Sixteen non-trunked simplex channels,[*] with the same 2% probability of blocking, would support an offered load of 16×0.02, or 0.32 E.[†] For these assumed conditions, trunking yields about an eight-fold increase in the amount of offered load, versus an equivalent number of non-trunked radio channels.

To equate the offered traffic load to the number of talk groups that can be carried by a trunking system, the average duration of a conversation, known as circuit holding time H, the average call request rate μ, and the maximum permissible queuing time t_q must be known. The product of the first two terms gives the offered traffic load per talk group, A_u. The total number of talk groups U is the quotient of the total offered traffic load A (from Figure 24) and the offered traffic load per talk group A_u. That is,

$$A_u = \mu H$$

$$U = A / A_u \tag{27}$$

Figure 24 gives the probability that a channel request will be delayed. Multiplying this probability by the factor $e^{-\frac{(C-A)t_q}{H}}$ gives the probability that a channel request will be delayed longer than time t_q, where C is the number of traffic channels and the other variables are as previously defined. After time t_q, the delayed call is dropped.

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[*] The comparison of seven trunked duplex channels to sixteen non-trunked simplex channels is made because: (1) repeaters require one channel for transmit and a second channel for receive, while point-to-point simplex uses the same channel for both transmit and receive, and (2) a trunked system will dedicate one channel-pair to the control channel so it will not be available as a traffic channel.

[†] The offered radio traffic load, in Erlangs, of a single non-trunked radio channel is equal to the probability, in percent, that a user will not gain immediate access to a traffic channel, divided by 100.

8 EQUIPMENT COSTS

A market survey of TIA-102 radios yielded six* major suppliers. Their product offerings are summarized below in Table 4. More products, from these and new manufacturers not listed here, are expected to enter the marketplace soon, but as of this writing, they are still awaiting FCC type acceptance certification.

Table 4. Project 25 Radio Prices

Cost →	Voice + Data			DES-OFB	Project 25	Pgm/Maint	O&M manuals	Fact/Vendor	Req'd Misc.	Subtotal
Radio ↓	Portable	Mobile	Base/Rptr	Encryption	Trunking	Software		Training	Accessories	
Manufacturer "A"	2075			150		500	125	400	405	3655
Manufacturer "B"	1710			495		135	150		611	3101
Manufacturer "C"		1850		495		135	150		113	2743
Manufacturer "C"			8606	495		170	225			9496
Manufacturer "D"	1439					1061	92		973	3565
Manufacturer "D"		1507				1021	86		711	3325
Manufacturer "D"			4887			898	132	1200	1958	9075
Manufacturer "E"	2180					297	148		220	2845
Manufacturer "E"			15157			297	148	370		15972

Collocated repeaters at base station sites require RF combining equipment as was discussed in Chapter 5. Representative equipment prices are given in Table 5.†

Table 5. RF Combiner Prices

Combiner Manufacturer	Configuration	Design	Tx System Loss (dB)	Rx System Loss (dB)	System Cost	Cost per Channel
Manufacturer #1	T8/R8 w/2 Antennas	Cavity	7.0	8.0	$64K	$8.0K
	T8/R8 w/1 Antenna	Cavity	8.0	8.0	$75K	$9.4K
	T5/R5 w/2 Antennas	Cavity	4.5	4.5	$33K	$6.6K
	T5/R5 w/1 Antenna	Cavity	6.0	6.0	$47K	$9.5K
	T4/R4 w/2 Ant (Var 1)	Cavity	4.0	4.5	$19K	$4.7K
	T4/R4 w/2 Ant (Var 2)	Cavity	5.5	7.0	$28K	$7.0K
	T4/R4 w/1 Ant (Var 2)	Cavity	7.0	7.0	$38K	$9.5K
Manufacturer #2	T8/R8 w/2 Antennas	Hybrid	12.5	1.0	$17K	$2.1K
	T8/R8 w/1 Antenna	Hybrid	13.5	1.0	$20K	$2.5K
	T5/R5 w/2 Antennas	Hybrid	9.0	1.0	$12K	$2.3K
	T5/R5 w/1 Antenna	Hybrid	10.0	1.0	$15K	$3.0K
	T4/R4 w/2 Ant (Var 1)	Hybrid/Cavity	8.0	1.0	$9K	$2.3K
	T4/R4 w/2 Ant (Var 2)	Hybrid	8.0	1.0	$9K	$2.3K
	T4/R4 w/1 Ant (Var 2)	Hybrid	9.0	1.0	$12K	$3.0K
Manufacturer #3	T8/R8 w/2 Antennas	Cavity	3.8	1.0	$16K	$2.0K
	T8/R8 w/1 Antenna	Cavity	6.0	1.0	$19K	$2.5K
	T5/R5 w/2 Antennas	Cavity	3.5	1.0	$10K	$1.2K
	T5/R5 w/1 Antenna	Cavity	5.2	1.0	$13K	$1.6K
	T4/R4 w/2 Ant (Var 1)	Cavity	3.9	1.0	$10K	$1.2K
	T4/R4 w/2 Ant (Var 2)	Cavity	3.9	1.0	$10K	$1.3K
	T4/R4 w/1 Ant (Var 2)	Cavity	5.2	1.0	$12K	$1.5K
Manufacturer #4	T8/R8 w/2 Antennas	Cavity	4.9	4.1	$18K	$2.3K
	T5/R5 w/2 Antennas	Cavity	4.3	3.5	$11K	$2.3K
	T4/R4 w/2 Ant (Var 1)	Cavity	4.1	3.3	$9K	$2.3K
	T4/R4 w/2 Ant (Var 2)	Cavity	4.5	3.7	$9K	$2.3K
Manufacturer #5	T8/R8 w/2 Antennas	Hybrid/Cavity	6.0	3.0	$14K	$1.8K
	T8/R8 w/1 Antenna	Hybrid/Cavity	10.0	7.0	$20K	$2.5K
	T5/R5 w/2 Antennas	Cavity	3.3	3.0	$9K	$1.8K
	T5/R5 w/1 Antenna	Cavity	7.3	7.0	$15K	$3.1K
	T4/R4 w/2 Ant (Var 1)	Cavity	2.8	2.3	$8K	$1.9K
	T4/R4 w/2 Ant (Var 2)	Hybrid/Cavity	6.0	2.5	$9K	$2.1K
	T4/R4 w/1 Ant (Var 2)	Hybrid/Cavity	10.0	6.5	$15K	$3.7K

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* One of the suppliers manufactures base station equipment for resale under license by other portable/mobile equipment manufacturers and therefore is not explicitly identified in Table 4.

† Private communication, Motorola (Salak), 1998. By way of example, in “Configuration,” “T5/R5 w/1 Ant” means a 5-channel repeater duplexed onto a single antenna, while “w/2 antennas” means separate antennas for transmit and receive. “(Var 1)” and (Var 2)” refer to proposed variations in the 4-channel combining schemes.

Antenna costs, both mobile and base, are neglected here. Likewise, costs for towers, communications huts, coaxial cable, etc., are not considered here because it is assumed that existing antennas and facilities will be reutilized to the maximum extent possible as the present analog FM systems are replaced.

9 AN ALTERNATIVE BAND PLAN

An alternative to the WCTF band plan is discussed in this chapter. This alternative plan, hereinafter referred to as the "ITS band plan," occupies the same block of railroad VHF spectrum as does the WCTF band plan. That is, both plans are composed of 181 channels spaced every 7.5 kHz, from 160.215 MHz to 161.565 MHz.

9.1 Description

A graphic presentation of the ITS band plan is shown in Figure 25. Compare this band plan to the WCTF band plan shown in Figure 1. The structure of the two band plans is very similar. The ITS version has two blocks of spectrum, from 160.215 to 160.7925 MHz and 160.950 to 161.5275 MHz, that comprise 78 trunked duplex voice channel-pairs. The two blocks of spectrum comprised of 5 channels apiece (from 160.800 to 160.830 MHz and 161.535 to 161.565 MHz) are non-trunked duplex channel-pairs. The remaining block of spectrum consists of 15 simplex channels (from 160.8375 to 160.9245 MHz).

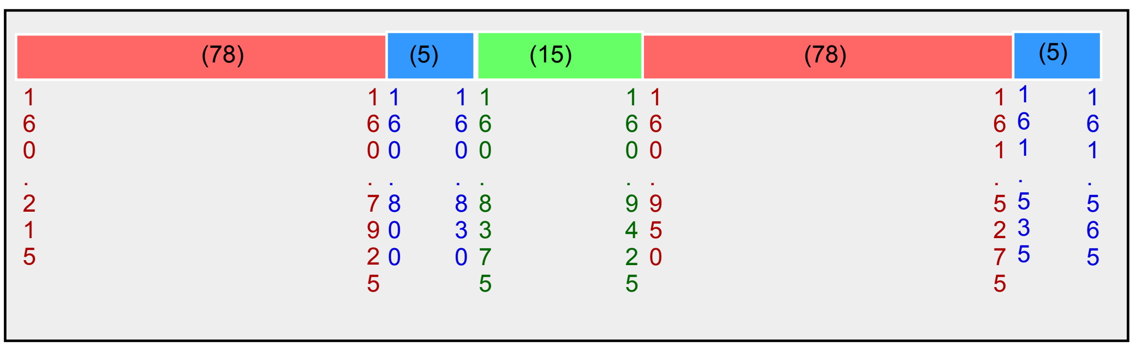

Figure 25. Railroad VHF band and the alternative ITS band plan.

The 78 duplex channel-pairs are apportioned among thirteen trunk groups of six channel-pairs each. This is in contrast to WCTF's ten trunk groups of eight channel-pairs each. The trunk groups of the ITS band plan are interstitially placed in the frequency domain, as depicted in Figure 26. Compare this figure to Figure 2.

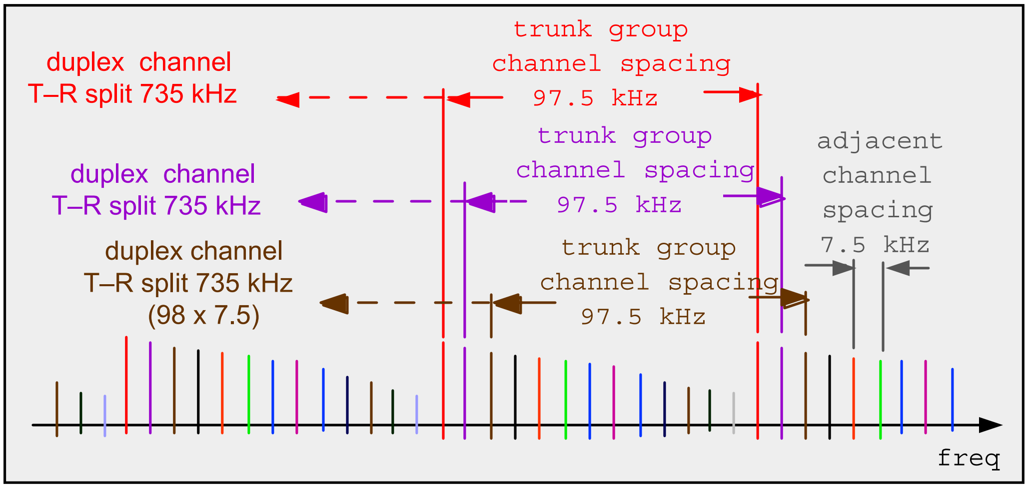

Figure 26. ITS band plan's channel spacing.

The channel spacing at each site, and the transmit-receive frequency split differ between the two plans, but both versions will easily support legacy analog FM systems, as the frequency splits of both plans are multiples of 15 kHz.*

During the migration period from legacy analog FM to TIA-102 digital, the arrangement of interstitial channels is somewhat different between the two band plans, but they are essentially equivalent. For the WCTF plan, legacy systems that cannot be programmed to the new interstitial channels (i.e., at odd multiples of 7.5 kHz, such as 160.7925 MHz) are constrained to operate only on half of the available trunk groups. The frequencies of these trunk groups are divisible by 15 kHz. Under the ITS band plan, older legacy analog FM radios that cannot be channeled to these interstitial frequencies are constrained to operate on half of the frequencies in each one of the trunk groups.†

* Transmit-receive frequency splits — WCTF: 720 kHz. ITS: 735 kHz.

† For example, frequency assignments of two adjacent trunk groups for both the WCTF and ITS band plans are:

WCTF Band Plan				ITS Band Plan			
Trunk Group A		Trunk Group B		Trunk Group A		Trunk Group B	
RX	TX	RX	TX	RX	TX	RX	TX
160.2150	160.9350	160.2225	160.9425	160.2150	160.9500	160.2225	160.9575
160.2900	161.0100	160.2975	161.0175	160.3125	161.0475	160.3200	161.0550
160.3650	161.0850	160.3725	161.0925	160.4100	161.1450	160.4175	161.1525
160.4400	161.1600	160.4475	161.1675	160.5075	161.2425	160.5150	161.2500
160.5150	161.2350	160.5225	161.2425	160.6050	161.3400	160.6125	161.3475
160.5900	161.3100	160.5975	161.3175	160.7025	161.4375	160.7100	161.4450
160.6650	161.3850	160.6725	161.3925	-	-	-	-
160.7400	161.4600	160.7475	161.4675	-	-	-	-

For the WCTF band plan, all frequencies from Trunk Group B (and similarly for Trunk Groups D, F, …) are from the new interstitial frequencies. For the ITS band plan, half of the frequency-pairs in each trunk group are from the new interstitial frequencies.

The non-trunked channels could be configured as a single 5-channel-wide duplex data channel and as a single 15-channel-wide simplex data channel, or as a single asymmetric duplex data channel.*

9.2 Multichannel Passband and Guardband Requirements

Figure 27 shows the aggregate transmission bandwidth of a multichannel RF signal utilizing ITS's band plan, along with the corresponding "guardband" between the most closely-spaced transmit and receive frequencies. Also shown in the figure, as dashed lines, are the WCTF passband/guardband data from Figure 11. An individual site complying with the ITS band plan† and using all six possible channels would require a transmission bandwidth of $(6 - 1) \times 97.5 + 12.5 = 500$ kHz. With a transmit-receive frequency split of 735 kHz on a single channel, the separation between the lowest-frequency transmit channel and the highest-frequency receive channel is $735 - 500 = 235$ kHz.

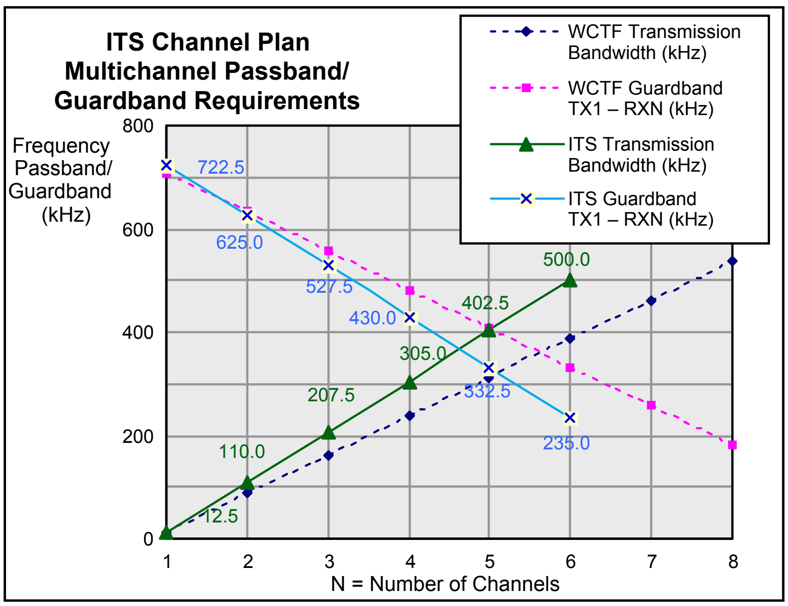

Figure 27. ITS band plan passband and guardband constraints.

Figure 27 can be somewhat misleading in that it appears that the WCTF band plan offers a narrower transmission bandwidth and a larger transmit-receive guardband for the same number of channels. This is true, but one must not forget that if a WCTF-configured site

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* Wider bandwidth channels permit faster data throughput. Duplex channels improve the efficiency of data transfer by reducing channel contention. Asymmetric channels permit a higher data throughput (using a larger bandwidth) in one direction than in the other direction.

† 735-kHz TX-RX frequency split, 97.5-kHz channel spacing.

is only using six or fewer duplex channels, then there are two or more channels at that site that are never used. This is a waste of radio spectrum resources. An alternative comparison between the two plans contrasts the passbands and guardbands of fully-utilized sites (i.e., the maximum number of channels used at each site), or of equal-percentage utilization (i.e., 50% = 3 channels for the ITS plan and 50% = 4 channels for the WCTF plan). Viewed in this light, the ITS band plan provides a narrower aggregate multichannel transmission bandwidth and larger guardband protection between the transmit and receive frequencies.

9.3 Intermodulation Interference

The method used by ITS to determine potential intermodulation frequencies was described in Section 5.5. To briefly recap, the bandwidth of a "k^{th}-order" harmonic or of a "k^{th}-order" intermodulation product frequency is "k" times the bandwidth of the original signals. Therefore, if the difference between the frequency of an intermodulation interference signal and the center frequency of a receiver is less than the sum of half the receiver bandwidth and "k" times half the transmitter bandwidth, that situation is "flagged" as an intermodulation "hit" (refer to Figure 13 and Equation (2)).

Using this methodology, an intermodulation analysis was conducted on both the WCTF and ITS band plans. Neither plan had any third-order hits. However, the WCTF plan showed potential for fifth-order intermodulation interference, whereas the ITS plan showed no intermodulation potential until the seventh-order.*

Section 5.5 noted that while the above approach identifies the *potential* for intermodulation interference to occur, whether it actually does or not depends on many other factors such as:

- individual transmitter power spectral densities,
- ambient receiver noise level,
- intermodulation rejection specification of the receiver,
- amount of attenuation that transmitter signals experience as they propagate through the various hybrids, cavities, postselectors, and preselectors in the transmitter-to-receiver path, and
- the probability that the offending transmitters are simultaneously transmitting.

Nevertheless, the ITS band plan indicates less potential for the occurrence of intermodulation interference than does the WCTF band plan, since seventh-order IMLs will be at lower power levels than fifth-order IMLs.

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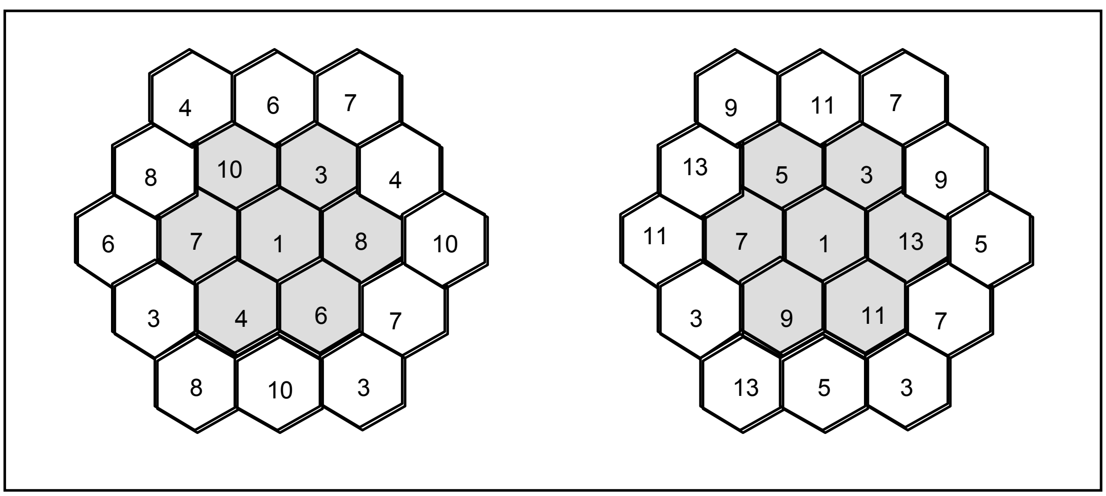
* Data output analysis files are available in electronic format from the author.

9.4 Trunking and Traffic Load Capacity

Erlang traffic theory was discussed in Chapter 7. Figure 24 showed that WCTF's eight-channel trunked design is clearly superior to ITS' six-channel design insofar as traffic handling capacity is concerned. For a 2% probability of channel requests being queued, the WCTF plan has about 67% greater traffic capacity than the ITS plan. However, the WCTF band plan is more limited in its possible cellular topologies.

9.5 Cellular Topologies

Section 4.2 notes that TIA-102 radios have a bandwidth of 12.5 kHz, but the railroad spectrum is channeled every 7.5 kHz. This means adjacent-channel transmissions from neighboring cells might interfere with intended reception of on-channel signals. One way to circumvent this problem is to assign the frequencies of the neighboring cell sites such that each site is at least two channels offset from all its neighbors. In a 7-cell topology, however, the WCTF band plan cannot achieve a 2-channel offset between all neighboring cells, whereas the ITS band plan can. These situations are depicted in Figure 28. Notice that after several iterations of swapping trunk groups around, the WCTF plan still has neighboring cells whose channel assignments are only offset by one channel (here, cells 3 and 4, and 6 and 7). With thirteen trunk groups available, the ITS plan can maintain a 2-channel offset between any one cell and all its neighbors. Until 6.25-kHz bandwidth equipment is available, the ITS plan provides the advantage in adapting to different cellular topologies.



(a) WCTF (b) ITS

Figure 28. Implementing a 7-cell topology with neighboring cells offset in frequency by two channels (15 kHz).

Section 9.1 noted that some older analog FM systems cannot operate on the new interstitial channels. This would limit the WCTF band plan to half of the ten trunk groups during the migration period. In this situation, a 7-cell topology cannot be implemented. With the ITS band plan, the new interstitial frequencies are commingled with the "legacy" frequencies in each and every trunk group, as the footnote on page 41 explains. What this means is that while the ITS band plan can accommodate 7-, 9-, 12-, and 13-cell topologies, each cell has only 2, vice 5, traffic channels available. Using the example scenario described in Chapter 7 (2% probability of a channel request being queued), Figure 24 indicates that a trunked 3-channel ITS-configured cell site can only carry 0.21 E of radio traffic. In contrast, a 6-channel site can carry 1.5 E. This is still an improvement, however, over the 0.06 E that a non-trunked 3-channel duplex site could carry, and is comparable to the 0.32 E that sixteen non-trunked simplex channels could carry.

9.6 Comparison of the ITS and WCTF Band Plans

Both the WCTF and ITS band plans have their own advantages and disadvantages. Table 6 summarizes the salient features of both plans. Reviewing these features in the context of all the previous discussions should enable the railroads to select a band plan best suited to their needs.

Table 6. Features of the ITS and WCTF Band Plans

Parameter	WCTF	ITS
Spectrum	160.215-161.565	160.215-161.565
Total no. of Channels	181	181
No. of Trunked Duplex Channels	80	78
No. of Trunk Groups	10	13
No. of Control Channels per Trunk	1	1
No. of Traffic Channels per Trunk	7	5
Traffic Cap'y @ 2% Blocking	2.6 E	1.5 E
No. of Non-Trunked Duplex Channels	5	5
No. of Simplex Channels	11	15
Transmitter Site Channel Spacing	75 kHz	97.5 kHz
Repeater Frequency Split	720 kHz	735 kHz
Support 7-Cell Topology with 15-kHz Channel Spacings?	No	Yes
Lowest Order of Intermodulation Interference that could occur within f_{rx} ± 6.25 kHz	5th-order	7th-order

10 SUMMARY, ISSUES, AND OBSERVATIONS

The Federal Communications Commission (FCC) has encouraged LMR manufacturers to design new equipment that incorporates narrowband and trunking technologies in order to increase radio system capacity and improve spectrum utilization, particularly in the crowded very high frequency (VHF) bands [1]. New equipment designs that do not use narrowband techniques will likely be denied FCC equipment type acceptance. This directive does not affect existing equipment or newly manufactured equipment designed in accordance with existing type acceptances. However, as the state-of-the-art in radio technology continues to advance, newly introduced radio systems and equipment will, by necessity, be narrowband. Users should begin planning now to deal with migration and compatibility issues when they introduce new equipment and systems into their existing LMR infrastructures.

The railroads have responded to the FCC's directive by forming an ad-hoc industry working group, the WCTF, to address the challenge and opportunities offered by the FCC's realignment initiative. WCTF has proposed a channel plan and is developing an equipment migration strategy.

As the railroad industry has proceeded with this initiative, several issues have been identified. An assessment of each is proffered by the Institute for Telecommunication Sciences.

ISSUE: Performance of new equipment — within the constraints of the given channel assignments, will the new equipment perform adequately? Is it backward-compatible with the older equipment now in service? Is the new equipment a proprietary design, or does it conform to recognized industry standards, ensuring interoperability between different vendors' equipment?

ASSESSMENT: New TIA-102 "Project 25" digital equipment is backward-compatible with 15-kHz analog FM equipment. Project 25 is an open standard architecture, so manufacturers building equipment complying to the specification can interoperate with other Project 25 equipment. Project 25 is not a proprietary design. WCTF has recommended Project 25 as the railroad's standard VHF LMR platform.

ISSUE: Channel assignments — are there enough radio channels to support all the business processes of the railroad industry? Is the voice traffic loading equally distributed across the available channels? Are there any techniques that can be employed to increase the traffic capacity?

ASSESSMENT: A collection of radio channels, each dedicated solely to a specific class of users, does not lend itself well to distributing the offered message traffic load efficiently among the available spectrum. While some channels might be lightly loaded, other radio channels, such as dispatch, may be very heavily loaded, giving the impression of not enough spectrum being available to adequately support the railroads' business needs and processes.

Trunking increases the traffic capacity of finite communications resources by distributing the offered message traffic load equally among the available radio channels, so one radio channel is not heavily loaded while others remain virtually unused. The ability to assign system access priorities to talk groups helps ensure that critical functions can always gain access to the radio system by pre-empting routine (i.e., non-safety-critical) radio communications, should all available radio channels be occupied. Of the two band plans presented here, the WCTF plan provides about 67% greater trunked-traffic capacity than the ITS band plan; however, there could be situations where adjacent-channel interference could limit the usefulness of the WCTF band plan.

ISSUE: Interference — will the proposed new equipment cause interference to the older existing equipment? Will the older systems cause interference to the new systems? If repeaters and base stations will be collocated (such as would be the case for trunked systems), what is the likelihood that intermodulation interference will be present? If present, can it be adequately suppressed? Will cochannel and adjacent-channel interference be a problem?

ASSESSMENT: Two of the more troublesome issues that must be considered are adjacent-channel interference, and intermodulation interference arising from the collocation of multiple channels and their frequency assignments.

Current Project 25 radios are designed with a 12.5-kHz bandwidth, but railroad VHF radio channels are spaced every 7.5 kHz. This means adjacent-channel transmissions will encroach on desired signals. One approach to alleviate this situation is to ensure that transmitters in neighboring regions are offset in frequency by at least two channels (15 kHz). The WCTF plan presents more difficulties in implementing a typical 7-cell topology than does the ITS plan.

Using the methodology described in Section 5.5, the ITS band plan is more resistant to co-site intermodulation interference than is the WCTF band plan.

ISSUE: Coverage — For a given level of speech intelligibility or effective data throughput, will the proposed radio system provide adequate geographic coverage?

ASSESSMENT: Measurements performed on a "typical" Project 25 radio indicate that for a given level of speech intelligibility, the Project 25 digital modulation format is superior to the traditional analog FM format, providing greater geographic coverage.

11 REFERENCES

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