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RADIO ENGINEERING
MICROWAVE RADIO
NOISE TESTS FOR HEAVILY LOADED SHORT HAUL MESSAGE SYSTEMS

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1. GENERAL

1.01 This practice outlines and coordinates methods for evaluating and optimizing the noise performance of message loaded short haul microwave radio systems. Primary emphasis is placed upon the TL-2 and TM-1 radio systems, although nearly all of the material discussed may be adapted for use with other systems.

1.02 Tests and alignment procedures presented in this section are intended for use in analyzing performance variations when the normal installation and maintenance techniques prove inadequate. Although these methods are most useful for systems where the load exceeds 240 message circuits, it is desirable to check out newly installed systems under conditions of full load even though the initial service load may be light in order to eliminate trouble sources which would prevent later growth.

1.03 The primary tool used in this practice for system evaluation is the noise loading measurement, which is a comprehensive measure of the performance of a system carrying message circuits. Secondary measurements are discussed in this practice but their basic purpose is to improve noise loading performance. These secondary tests are required to locate trouble sources which cannot be found using standard routine maintenance techniques and test equipment.

2. SYSTEM CAPABILITY

2.01 The TL-2 and TM-1 systems are intended to provide a basic 250-mile system of ten radio hops. They are designed to carry 600 message circuits with a 10-hop worst-channel noise objective of no greater than 35 dbrnc0 (29 dba0) in the busy hour.

2.02 Allocating equal noise to each radio hop and assuming power addition of noise, the per-hop noise objective is 25 dbrnc0 (19 dba0). If in an actual installation the received signal level is different from the nominal −42 dbm minimum (−45 dbm after aging etc), the per-hop performance actually obtained will vary from the 25-dbrnc0 objective.

2.03 Pre-emphasis is usually required on TL-2 and TM-1 systems to meet the 600-channel objectives. When noise-loading tests are performed without pre-emphasis, the allocated noise objectives in the top and bottom channels are modified as follows:

(a) Bottom channel noise (70 kc): 19.5 dbrnc0 (13.5 dba0).

(b) Top channel noise (2438 kc): 27.5 dbrnc0 (21.5 dba0).

Note that the top channel noise-loading objective is 1-db low to compensate for the measuring frequency (2428 kc) being below the actual top frequency.

2.04 The purpose of this and the other heavily-loaded system practices is to prescribe tests and adjustments to obtain optimum performance from each radio hop within the capability of the equipment as it is utilized. This does not mean that 25 dbrnc0 or better will necessarily be realized on each radio hop, but when such performance is not obtained, the tests will indicate the reason.

3. TESTING OUTLINE

3.01 The tests referred to and analyzed in this practice may be performed in a variety of ways depending upon individual conditions and the available test equipment. Emphasis in this practice is placed upon an over-all approach to optimizing system performance, and the analysis of test measurements made to accomplish this end. Therefore, the actual detailed test procedures are covered in separate practices. Where possible, two or more methods for the same test have been described to maintain flexibility. These practices are to be found in the following series.

Section 409-403-504 — Idle Noise Measurement

Section 409-403-505 — Noise Loading Measurement

Section 409-403-506 — Envelope Delay Measurement

3.02 Transmission degradation is indicated by high telephone circuit noise, tones, crosstalk, or similar service impairments. In most cases, high-noise levels may be traced to a single-radio hop which may then be tested thoroughly on an out-of-service basis. Therefore, the test procedures emphasize single-hop measurements and analysis. The method for aligning individual hops that are to be heavily loaded is essentially the same as for lightly-loaded hops except that some additional testing and analysis will be required.

3.03 The initial establishment of radio transmission path is covered in Section 409-403-500. This section includes antenna orientation techniques as well as appropriate methods for determining RF signal levels. Initial alignment for the transmitter and receiver gains and frequencies are discussed in Sections 409-404-500 and 409-406-500 series. Once the hop is aligned according to these instructions, it will be possible to proceed with the tests discussed below. The testing procedure can be outlined as follows:

(a) Idle Noise Measurement

(b) Bottom Channel Noise Loading

(1) Linearization

(2) Evaluation

(c) Top Channel Noise Loading

(1) Evaluation

(d) Placing in Service

These tests are briefly described in Part 4 of this practice in an order which will minimize repetition. The detailed analysis of the test data and trouble clearing are treated in Parts 5 and 6.

4. TESTING PROCEDURE

4.01 *Idle Noise Measurement:* Once the radio path is established and the receiver is properly locked on the transmitted signal frequency, terminate the transmitter baseband input jack at the other end of the hop and measure the baseband noise out of the RCVR OUT jack of the receiver IF and baseband unit. Use a selective analyzer to measure the idle circuit noise as a function of frequency. The test procedure is described in Section 409-403-504. Refer to 5.01 of this practice for the analysis of idle noise measurements.

4.02 ***Bottom Channel Noise Loading***

(a) Prior to making the actual noise-loading run in the bottom channel of a new radio hop of one in which either the transmitter control unit or the transmitter klystron has just been replaced, it is necessary to linearize the hop and adjust the deviation sensitivity. If the hop has not been linearized previously, refer to Section 409-404-501 and follow the procedures for lightly loaded and nondiversity systems. Next, prepare to make a noise loading measurement at the bottom channel frequency (typically 70 kc) as given in Section 409-403-505. Use a drive level of −21 dbm into the transmitter baseband input jack. With the bottom channel slot or notch filter in the circuit, adjust the transmitter repeller voltage XMTR RPLR slowly for minimum bottom channel noise as seen at the receiving or far end of the hop. This minimum will indicate the optimum linearity point since modulation noise in the bottom channel is predominantly due to baseband nonlinearity. The transmitter should then be placed on frequency by means of the klystron cavity adjustment. The baseband drive or input signal is to be removed for this last adjustment. Failure to remove the input signal will degrade the accuracy of frequency setting.

(b) Check to see if the bottom channel noise with −21 dbm drive is within 1 db of the noise measured when the transmitter baseband input is terminated and has no drive. If this condition is not obtained, repeat the minimizing technique until the lowest noise level is obtained. If for some reason the bottom channel modulation noise remains high, continue with the rest of the noise loading run and refer to 5.03 of this practice for further suggestions.

(c) The deviation sensitivity of the transmitting klystron is slightly affected by changing the repeller voltage to obtain optimum linearity. It may therefore be necessary, after linearizing, to reset the transmitter deviation sensitivity as shown in Section 409-404-502.

Note: If the hop gain is found to be very high (approximately 6 db above normal), the receiver may accidentally have been aligned with a beat oscillator offset of 35 mc instead of 70 mc. Refer to Section 409-406-502 for correct alignment of the beat oscillator.

(d) Make a noise loading run versus drive level in the bottom frequency slot. Refer to Section 409-403-505 for the test methods. Determine from the results whether adequate performance and linearization have been obtained. Refer to 5.03 of this practice for analysis

4.03 ***Top Channel Noise Loading:*** If the tests are being performed sequentially, as suggested, it will not be necessary to make any additional adjustments before making the top channel noise-loading run. Rearrange the noise-loading equipment so that top channel measurements can be made and then make a noise-loading run as described in Section 409-403-505. The top channel frequency for 600-circuit loading is 2438 kc on some noise-loading sets, but other nearby frequencies are acceptable. Determine from the results whether adequate performance has been obtained at the reference drive of −24 dbm. Refer to 5.04 of this practice for analysis of this curve.

4.04 ***Placing Hop in Service:*** When the tests are completed and satisfactory results are obtained, the radio hop may be placed in service per Section 409-402-501.

5. ANALYSIS OF RESULTS

5.01 This part analyzes in detail the test results obtained from Part 4. In addition, where trouble is encountered, further tests are specified to assist in fault location.

5.02 *Idle Noise*

(a) The noise level measured as a function of frequency at the receiver baseband output jack with no signal applied to the transmitter baseband input is called idle noise. This noise is primarily the result of two sources, klystron shot noise and first-circuit or front-end noise. The shot noise is a function of the klystrons alone and does not vary with signal level or baseband frequency. The curve in Fig. 1 shows the typical shot noise in TL-2 to be approximately −98 dbm and TM-1 to be approximately −92.5 dbm, as measured on a J64037B Transmission Measuring Set (37B TMS) at the receiver baseband output jack. These values may be expected to vary ±1 or 2 db depending upon individual klystron characteristics. The thermal or front-end noise component varies almost db for db with received signal level and rises with baseband frequency at a rate of 6 db per octave.

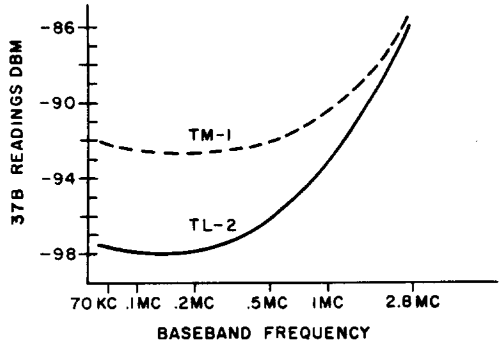

Fig. 1 – Typical Idle Noise Curves

(b) The thermal noise component of the total idle noise as measured on the 37B TMS may be described as follows:

$$\text{Thermal Noise} = [-138.5 + NF - P_r] + 20 \log \left(\frac{f}{4.0}\right) \text{ dbm}$$

Where NF is the receiver noise figure measured at the modulator input,

where P_r is the received RF signal power in dbm at the receiver modulator, and where f is the baseband frequency of measurement in megacycles. This relationship is useful as a check of the received signal level power and noise figure of the receiver. The total idle noise is the power sum of the shot noise and the thermal noise. Note that powers expressed in dbm cannot be added algebraically but must first be expressed in watts or millwatts.

Example: On a particular TM-1 hop, the noise measured at 100 kc with the 37B TMS is −93 dbm and the noise measured at 2438 kc is −87 dbm. The received RF signal power measured per the procedures in Section 409-406-403 is −45 dbm. What is the noise figure of the receiver?

Solution: The noise measured at 100 kc should be primarily shot noise (see Fig. 1), so

$$\text{Shot noise at 100 kc} = -93 \text{ dbm} = 0.5 \times 10^{-9} \text{ milliwatts}$$

$$\text{Total idle noise at 2438 kc} = -87 \text{ dbm} = 2 \times 10^{-9} \text{ milliwatts}$$

$$\text{Shot noise at 2438 kc} = -93 \text{ dbm} = 0.5 \times 10^{-9} \text{ milliwatts (same as at 100 kc)}$$

$$\text{Thermal noise at 2438 kc} = 2 \times 10^{-9} - 0.5 \times 10^{-9} = 1.5 \times 10^{-9} \text{ milliwatts} = \text{difference}$$

$$= -88.2 \text{ dbm}$$

From the formula given above:

$$-88.2 = 138.5 + NF - (-45) + 20 \log \frac{2.438}{4}$$

$$-88.2 = 138.5 + NF + 45 - 4.3$$

$$NF = 9.6 \text{ db}$$

This demonstrates how the idle noise components add as well as give a typical receiver noise figure. Actual noise figures for TL-2 and TM-1 receivers range downward from a maximum of 11 db down to approximately 8 db.

5.03 ***Additional Sources of Idle Noise:*** Tones and other noise may appear during the idle noise measurement as the result of interfering signals or faulty components. Some of the sources are as follows:

(a) ***Tones at Approximately 2-kc Intervals Extending Upward in Frequency from Below 70 Kc:*** In some cases, tones may result from harmonics of the inverter frequencies used in the high-voltage supplies for TL-2, TM-1, or the TM-A1 power amplifier. These tones may be coupled in through common source impedances, poor cable dress, defective plug-in units, or in the case of TM-1, 200- and 400-volt regulator misadjustments. The generator of these tones can often be identified by slightly altering the KLYSTRON ADJ VOLTS control on each suspected inverter in turn. Since the inverter frequency is voltage sensitive, listening for a change in the pitch of the tone at the receiver baseband output will establish an association. If the tones are picked up by, or ahead of the transmitting baseband amplifier, they will disappear when the 2-ampere fuse is removed on the 20-volt regulator panel at the transmitting end. Of course, this will disable the receiver on the same panel and appropriate measures must first be taken to avoid the interruption of active circuits in an operating system. If faulty plug-in units have been eliminated as potential tone sources and the tones remain, the power wiring should be checked to verify that it is in accordance with SD-97158-01, SD-97264-01, or SD-97303-01.

(b) ***Isolated Clear Tones Between 70 Kc and 3 Mc:*** These tones are not caused by direct RF or IF carrier interference but will be caused by baseband pickup of signals from other equipment operating in the vicinity of the radio equipment. If the tones were introduced through RF or IF pickup, they would appear as smears of noise or crackling as heard on the 37B TMS at baseband frequencies rather than as clear tones [see (c)]. Sometimes the tones will not actually be located at the indicated frequency but will be caused by intermodulation products in the selective measuring equipment. In the latter case, it is helpful to use a low-pass filter ahead of the test set to block frequencies above 10 mc which may generate spurious products. A 544A filter may be used although it is flat only to 4 mc. Lacking a low-pass filter, the 37B TMS may be tested for spurious product generation by reducing the input level in 1-db steps with an external attenuator. If the measured tone drops by 2 db or more, an internal product in the 37B TMS is indicated.

(c) ***Noise Bumps or Crackling Noise in Approximately a 50-kc Bandwidth:*** This noise may be generated by an interfering signal either at radio frequency or intermediate frequency. This noise tends to jump around in frequency and is sometimes difficult to find and hold. An IF spectrum analyzer is very useful in locating the cause of this form of interference. If such an instrument is obtainable, observe the spectrum at the IF monitor jack of the receiver where the interference is first detected. If necessary for display reasons, reduce the incoming carrier level by partially inserting the transmitter waveguide switch so that the incoming carrier and front-end noise can be seen as shown in Fig. 2. The interfering tone will normally be found separated from the carrier frequency by the frequency of the noise bump in the baseband signal. In order to determine whether the interfering frequency is at radio frequency or intermediate frequency, first turn off the receiver AFC switch and slowly turn the receiver BO RPLR control from side to side approximately 1/4 turn while watching the display. If the frequency spacing between the desired carrier and the interfering tone remains constant, the interfering signal is approximately at the signal fre-

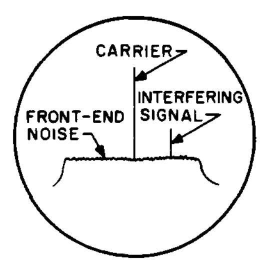

Fig. 2 – Spectrum Analyzer Display of Interference

quency. If the carrier and tone move at different rates, an image or higher order product is involved. The frequency difference between the desired signal and the interfering tone will be the baseband center frequency of the noise bump. If the interfering tone remains stationary on the display as the BO RPLR control is turned, the interfering signal is at intermediate frequency, or results from two radio frequency signals mixing with each other rather than with the beating oscillator. If the display behaves in a way which is different from above, the following suggestions are given.

(1) If the carrier and interfering signals move in opposite directions with the BO RPLR rotation, look for an interfering RF signal at the image frequency (example, carrier 10,755 mc, beat oscillator 10.825 mc, image 10,895 mc).

(2) Another possibility for this effect is a preamplifier oscillation at 140 mc which may be stopped by introducing 3- to 10-db attenuation directly at the preamplifier output. This oscillation is occasionally found in TM-1 preamplifiers and is a trouble condition requiring replacement of the unit.

(3) As an alternative test, leave the receiver AFC on and slowly offset the transmitter repeller control approximately 1/8 turn and allow the receiver AFC to follow. In this case, the baseband output frequency of the interference will change 1 to 2 mc if it is due to an RF interference but only 50 to 100 kc if it is at the intermediate frequency.

(d) The previous tests were based on the use of an IF spectrum analyzer. If this instrument is not available, a baseband analyzer may be used but will require more careful reasoning to identify the interference source. The method is similar, however, and the technique of shifting the beat oscillator frequency and observing the resulting interference shifts may be used but the interpretations will differ.

5.04 *Bottom Channel Noise Loading*

(a) Depending upon alignment and klystron noise, a typical bottom channel noise loading run versus drive for TL-2 will be similar to that shown in Fig. 3. The curve for TM-1 will be higher by approximately 4 db. The curve may be broken into two components, noise resulting from idle noise and noise resulting from modulation noise. The straight line portion of the curve at signal drive levels less than approximately −21 dbm is predominantly due to idle noise. As the baseband drive is increased, the contribution of idle noise in dbrnc0 will continue to decrease at a db-for-db rate, as indicated by the dotted line. When the baseband signal drive is increased above approximately −21 dbm, the total noise begins to increase. This increase is due to the increasing amount of modulation noise with increasing drive.

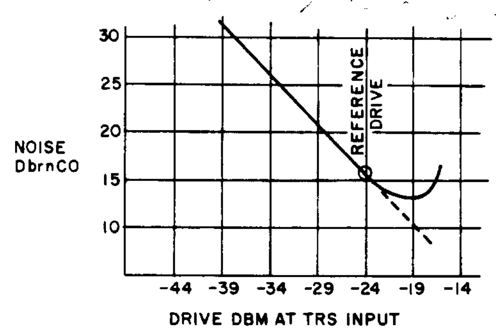

Fig. 3 – Noise Loading Curve, Bottom Channel

(b) The contribution of idle noise may be checked by noting the intercept point (marked by the circles in Fig. 3). The noise at the intercept point in the bottom channel (typically 70 kc) should correspond to the idle noise at the same frequency (70 kc in this example) measured previously. The relationship is given as follows (see Part 9 for derivation of this and other equations used in this practice):

$$\begin{bmatrix} \text{Bottom Channel Noise} \\ \text{Measured with the} \\ \text{37B TMS in db} \end{bmatrix} + 112 = \begin{bmatrix} \text{Bottom Channel Noise} \\ \text{Intercept in dbrnc0} \\ \text{(without pre-emphasis)} \end{bmatrix}$$

Example: Consider an idle noise measurement, using a 37B TMS, of −98 dbm.

$$[-98] + 112 = 14 \text{ dbrnc0 intercept}$$

If agreement within 2 db is not found between the intercept and its calculated value, check the test setup and radio hop for proper levels and the noise generator for a flat output spectrum within ±1 db as specified previously.

(c) Notice that the intercept point is defined as the contribution due to idle noise when the baseband drive is at the reference value of −24 dbm. Consider the case where the actual noise at the reference drive (−24 dbm) is due to both idle noise and intermodulation noise as in Fig. 4. The intercept point is still defined as the contribution due to idle noise (see Fig. 4 for example).

(d) The idle noise part of the bottom channel noise is nearly always controlled by klystron shot noise (shot noise comes from both the transmitting and receiving klystron nearly equally and is independent of received RF signal level). Excessive shot noise may usually be corrected by replacing a noisy klystron at one end of the hop or the other. On rare occasions, the baseband amplifiers may contribute to the low-frequency idle noise. In this case, the trouble may be corrected by replacing either the transmitting control unit or the receiver IF and baseband unit, whichever is at fault. The transmitter baseband amplifier may be noise loaded by itself by monitoring its output via the MON jacks on

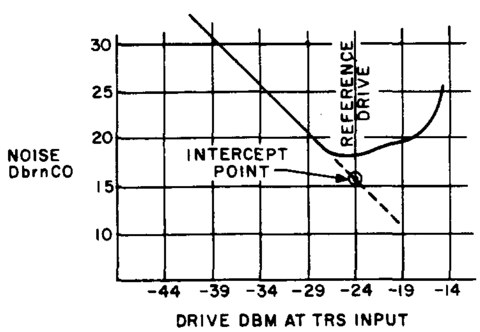

Fig. 4 – Noise Loading Curve Illustrating Intercept Point

the transmitter control unit. The resulting noise at the reference drive should be 12 dbrnc0 or less.

(e) Occasionally, a power supply harmonic tone may fall in the measuring slot causing an abnormally high-noise intercept. The tone may be identified using a 37B TMS with a headset. If the tone cannot be removed readily, it may be possible to move its frequency out of the slot as mentioned previously by adjusting the KLY ADJ VOLTS control on the appropriate inverter.

(f) The right-hand side of the bottom channel noise loading curve is controlled by modulation noise produced by circuits operating at baseband frequencies associated with the modulation and demodulation processes. Another way of stating this is to say the modulation noise in the bottom channel is predominantly controlled by the baseband linearity. In order of decreasing importance the nonlinearity contributors are: klystron repeller characteristic, receiver discriminator, and the transmitting and receiving baseband amplifiers. The linearizing technique using noise loading should yield a noise-loading result similar to that shown in Fig. 3. If difficulty exists in producing noise-loading curves where the modulation noise is similar to that shown in Fig. 5 (that is, not more than 1 or 2 db greater than shown), especially if there is indication of flattening of the curve bottom (see Fig. 4), poor hop linearity is indicated. The linearity may possibly be improved by first replacing the receiver IF and baseband unit. If no improvement results, the transmitter control unit should be replaced instead. It will be the exceptional case when replacement of one of these two items does not provide a linearity of sufficient quality to produce good noise-loading results. However, when that happens, it will be very helpful to be able to measure the linearity directly. Several test sets are available for doing this. One set that is suggested for portability and capability is the 10E-1MW and 70E2-MW manufactured by the Collins Radio Company. This set is also capable of measuring envelope delay, a valuable feature discussed later in this section. The actual measurement of linearity and delay is discussed in Section 409-403-506.

5.05 ***Top Channel Noise Loading***

(a) Depending on equipment alignment, a received signal level, and waveguide conditions, a top channel noise-loading curve may look like that shown in Fig. 5. As mentioned previously, the top channel on most noise-loading sets for 600-channel loading is 2438 kc.

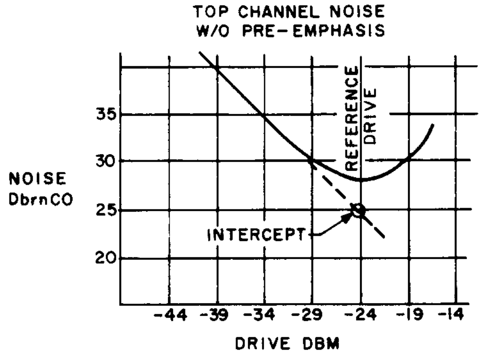

Fig. 5 – Top Channel Noise Loading Curve

(b) The character of the curve is similar to the noise-loading curve measured at the bottom channel in that the noise at the low baseband drive (less than approximately −24 dbm) is due primarily to idle noise and the noise for baseband drives above approximately −24 dbm will be primarily the result of modulation noise. The sources of idle and modulation noise are not the same as in the bottom channel, however.

(c) The idle noise in the top channel (left-hand side of the curve) is dominated by thermal or front-end noise. This subject has been discussed in detail in 5.01. As with the bottom channel, the top channel noise intercept is related to the noise measured with the 37B TMS by the following formula:

$$\left[\begin{array}{l}\text{Top Channel Noise}\\\text{Measured with the}\\\text{37B TMS in dbm}\end{array}\right] + 112 = \left[\begin{array}{l}\text{Top Channel Noise}\\\text{Intercept in dbrnc0}\\\text{(without pre-emphasis)}\end{array}\right]$$

Also as with the bottom channel, agreement between the calculated noise intercept as found above, and its measured value should be within 2 db if the test equipment is operating properly.

(d) The noise appearing on the right-hand side of the top channel noise-loading curve is generated primarily by modulation products due to envelope delay distortion. This form of distortion is the result of undesirable phase-versus-frequency shapes of the microwave and IF components in the transmission path. These characteristics are measured at the time of manufacture and should be within proper limits. However, one area of equipment which cannot be tested at the time of manufacture is the waveguide and antenna system.

(e) The majority of high top channel modulation noise troubles can be traced to reflections, resonances, or other discontinuities in the waveguide runs. The most useful piece of test equipment for measuring delay characteristics is called a delay set or a delay and linearity set. Instructions for using such a set are included in Section 409-403-506.

(f) An actual poor noise-loading curve resulting from multiple reflections in a 100-foot waveguide run is shown as "before" in Fig. 6. In this particular case, the delay measurement disclosed a delay ripple induced

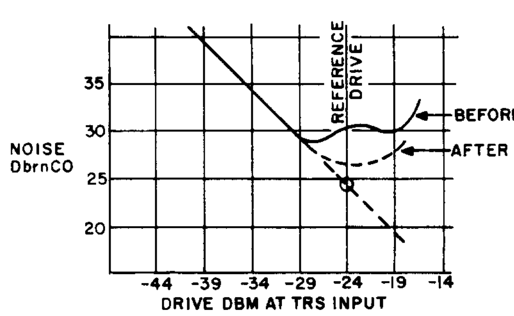

Fig. 6 – Top Channel Noise Loading Curve Showing Effect of Delay Ripple

by a section of flexible waveguide which was damaged at the time of installation. A list of sources of microwave delay distortion follows. This list is not exhaustive but includes the sources most often found.

(1) Deformed waveguide

(2) Coupling effects in antennas or waveguides

(3) Waveguide switch not seated

(4) Foreign object in waveguide

(5) Mistuned microwave networks

(6) Insufficient frequency spacing between adjacent channels on a waveguide run

A more detailed delay analysis based on the use of a delay measuring set is contained in Part 6.

6. DELAY TESTING

6.01 Delay, or more properly envelope delay, is defined as the rate of change of phase versus frequency in the RF or IF portions of the radio system. Phase variations in a radio system are caused by frequency-sensitive components in the transmission path whether intended or not. Linear phase variations versus frequency produces no modulation products in an FM system and the resulting delay or phase slope is constant. Parabolic phase versus frequency causes second order modulation noise in an FM system and results in a linear delay slope. Phase ripples result in higher order modulation noise of a peculiar nature described later in this practice and, in turn, results in a delay ripple. A delay measuring method then is useful in locating sources of modulation noise in the FM system.

6.02 As a practical matter, short haul radio systems may be tested for delay and linearity on a per-hop basis by a relativly simple test signal applied to the baseband input of the transmitter and recovered at the receiver baseband output. The use of delay measuring equipment is described in Section 409-403-506. The test signal consists of two parts: the first, a high-frequency sine wave (approximately 300 kc) of low amplitude which is transmitted over the system. This signal is recovered and measured for variations in phase and amplitude as a function of the second signal which is a low-frequency sine wave of large amplitude. The second signal in effect scans the band by moving the transmitter center frequency back and forth at a low rate while the first signal reads the delay and linearity on a nearly point-by-point basis. Variations in the received phase are detected and displayed as the delay pattern.

6.03 Ideally, the linearity and delay traces would be horizontal lines. In practice, the best obtainable delay picture with TL-2 or TM-1 looks like a broad and very much rounded W. The important contributor to second-order modulation noise in the radio hop is the slope of the delay curve at midband. The other curvature contributes primarily to higher-order modulation noise which is not normally controlling in the TL-2 or TM-1 systems.

6.04 If fine grain sinusoidal ripples are present in the delay pattern, a peculiar form of modulation noise may be found as is described in the section on noise loading (Part 5). This noise tends to flatten the bottom of the noise-loading curve as shown previously in Fig. 6.

6.05 In operation, the low-frequency output of the delay set is adjusted in amplitude to produce a peak-to-peak deviation of approximately 10 mc in the radio channel being tested.

6.06 The high-frequency signal level is normally set to the minimum value consistent with a stable display. Increasing the high-frequency level beyond this point tends to smooth out the delay pattern and obscure the kinks or ripples which are being sought.

6.07 A typical delay pattern is shown in Fig. 7 for a well adjusted TM-1 hop. Shown also are the two primary components which make up this pattern and which result from the IF and RF circuits.

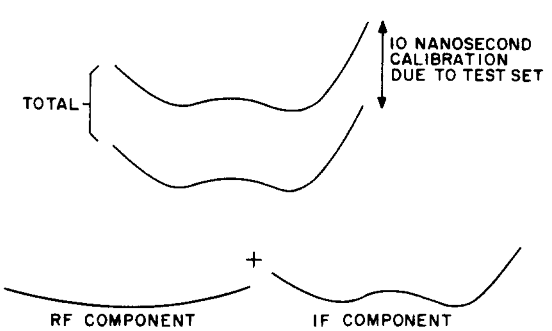

Fig. 7 — Delay Components

It is not possible, in the field, to display the IF and RF components separately but some tricks may be used to isolate the source of kinks or ripples when they occur. Take, for example, the case previously cited in which a noise-loading test turned up a radio hop with a flat-bottomed noise curve in the top channel. This condition is almost invariably associated with a rippled or sharply kinked delay pattern. The actual delay pattern which was found in this case is shown in Fig. 8.

Fig. 8 – Delay Patterns Showing Ripples

A simple first step in locating the source of anomalies was to swap the receiver IF and baseband unit with a spare. In this case, no appreciable change was found and the next step was to move the signal around in the IF band without changing the transmitter microwave signal frequency and then to reverse the process by moving the microwave signal around in frequency and holding the IF constant.

6.08 The procedure for shifting the microwave of RF delay and IF delay separately is as follows:

(a) To move the signal at intermediate frequency only, turn off the receiver AFC and rock the BO RPLR control slowly from side to side approximately 1/8 turn. This causes the delay components at the intermediate frequency to shift back and forth on the oscilloscope, but not the components due to microwave or radio frequency sources.

(b) To move the microwave or RF signal around, but not the IF, leave the AFC on and rock the TRS RPLR control very slowly approximately ±1/8 turn on the transmitter. In this way, the AFC system will keep the IF signal centered at 70 mc and only the delay components due to the microwave circuits will shift back and forth on the oscilloscope.

In the example cases, the delay ripples were found to move with changing transmitter frequency and, therefore, were associated with the microwave circuitry. The moving component was a sinusoid in delay with a peak-to-peak amplitude of 4 nanoseconds and a period of repetition rate of 1 cycle per 4 mc of frequency change. The frequency spacing was determined by means of an absorption IF meter (J68345J-90, List 2) placed between the receiver preamplifier and IF and baseband amplifier unit.

6.09 ***Delay Ripple Generation and Repetition Rates***

(a) Fig. 9 illustrates one means of sinusoidal delay ripple generation in a microwave system. The desired or direct signal from the radio transmitter is radiated from the antenna. In addition, a delayed signal is also radiated. The phase angle the delayed signal makes with the direct signal varies directly with the change in transmitter frequency. As a result, the envelope delay pattern varies sinusoidally with frequency. As the physical length of the reflection path in increased, the ripples will be spaced more closely. It is possible to determine the distance in feet between the two reflections causing a sinusoidal ripple by measuring the frequency difference between two adjacent ripple peaks. Table A lists conversion factors for this purpose.

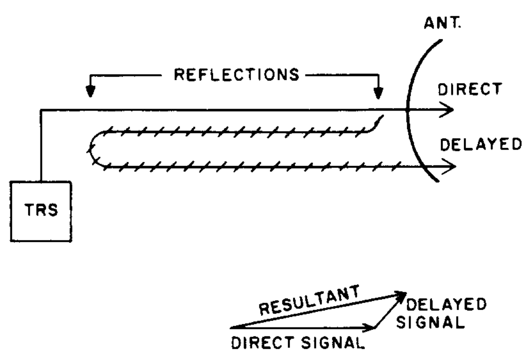

Fig. 9 – Multiple Reflections Causing Delay Ripple

TABLE A — RIPPLE SPACING DUE TO MULTIPLE REFLECTIONS									
WAVEGUIDE MEDIUM	MODE	FREQUENCY, GIGACYCLES							
		5.925	6.425	10.7	10.9	11.1	11.3	11.5	11.7
		CONVERSION FACTOR							
WR90	TE 10	—	—	388	393	397	401	404	407
WR159	TE 10	388	402	462	463	464	465	466	467
WR159	TE 20, 01	—	—	354	360	366	371	376	380
WR159	TE 30	—	—	—	—	—	82	122	150
Air	Any			491		All			

$$\text{Ft between reflections} = \frac{\text{Conversion Factor}}{\text{Spacing, mc}} = \frac{\text{Conversion Factor}}{\text{S}}$$

Example: WR159, 6425 mc, 4-mc ripple spacing

$$\frac{402}{4} = 100.5 \text{ feet between reflections}$$

(b) The delay ripple in the previous example was traced by this method to a reflection near the antenna end of a 100-foot waveguide run. The reflected signal returned to the transmitter from which it was re-reflected and radiated. After a defective flexible waveguide section was replaced, the noise loading curve marked "after" in the previous section was obtained (see Fig. 6).

6.10 A useful test for locating echoes of the type just illustrated is to insert a waveguide attenuator with approximately 3-db loss in the suspected waveguide run. The effect of such an attenuator is to reduce the direct or desired signal by 3 db and the delayed or echo signal by 9 db. The resulting 6-db change in desired-to-delayed signal ratio will reduce the amplitude of the ripples to one half. This effect will take place ***only*** if the attenuator is located between the two reflecting points. Therefore, moving the attenuator from one waveguide joint to another will help isolate a faulty component.

6.11 Another cause of delay ripple is illustrated in Fig. 10. In this case, attenuation put in the WR90 waveguide had no effect on the ripple amplitude. As may be seen in Fig. 10, the reflection path was totally outside of the 11-gigacycle waveguide system. A correction was made to include a low-pass filter directly at the edge of each dual-frequency antenna at a point approximately 8 feet removed from the antenna feed structure. It was later observed, however, that even with this change, certain radio channels would experience delay distortion while others would be unaffected.

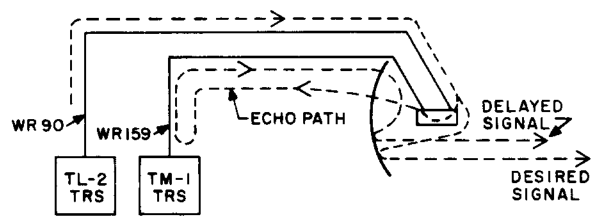

Fig. 10 — Delay Caused by Reflections in Another System

6.12 Further investigation using a delay set showed that the delay pattern had changed from a sinusoidal ripple to isolated bumps or pips in the pattern. The mechanism of generation of this delay pattern is shown in Fig. 11.

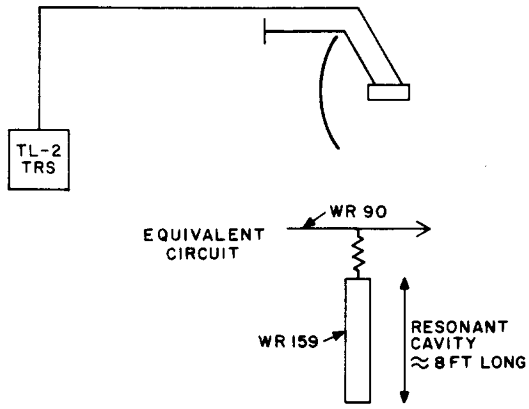

Fig. 11 – High-Q Resonance Which Causes Delay Bumps

The 8-foot section of the WR159 waveguide now acted as a high-Q resonant cavity with resonances spaced at approximately 60-mc intervals (incidentally this spacing agrees with the conversion factors listed in Table A for 11.5-gigacycle signals in WR159 waveguide

$$\text{that is: } \frac{466}{8} \approx 58\text{-mc spacing}).$$

The long term solution to this problem is a selective absorber to remove rather than reflect the higher-band signals rather than absorb them. This absorber is the KS-20148 filter.

6.13 Another waveguide system which produced excessive delay bumps is illustrated in Fig. 12. In this instance, a section of circular WC150 waveguide was used at 11 gigacycles to obtain low loss. The unused polarization was capped off with shorting plates which actually formed a cavity which was resonant at intervals of 8 mc and caused the delay bumps. Two alternative solutions to this problem are available, the first is to rotate one of the polarizing networks very carefully to minimize the coupling to the unused mode. This adjustment was made using a delay set as a detector. The preferred method is to remove the shorting plate and replace them with terminations.

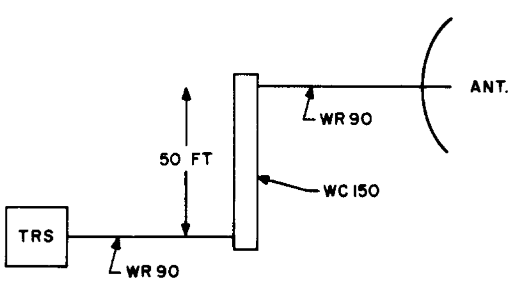

Fig. 12 – Circular Waveguide Causing Delay Bumps

6.14 A variety of delay ripple situations may be encountered in installations using the KS-15676 horn-reflector antenna and its associated WC281 circular waveguide run. One of

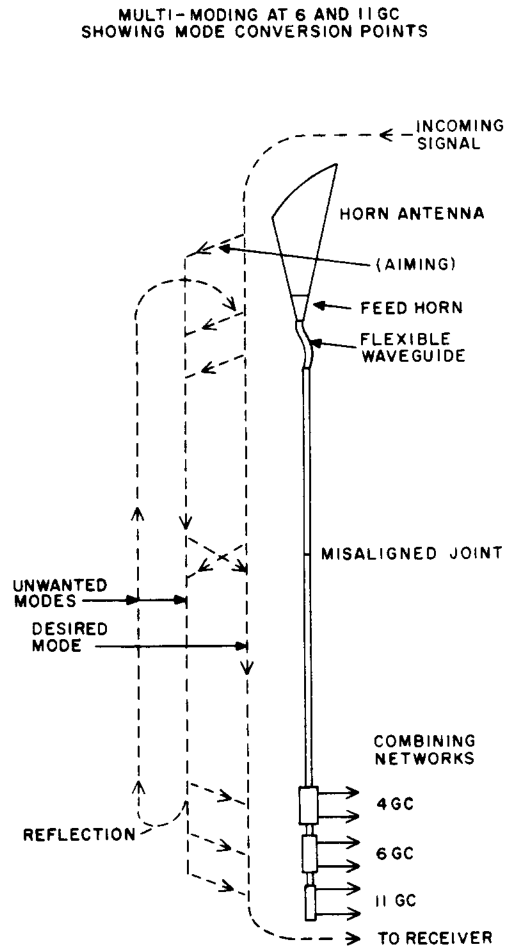

Fig. 13 – Multimoding and Reflections Using a Horn-Reflector Antenna

the forms of ripple may be generated by multiple reflections as shown previously. Another form, which is more difficult to correct, is referred to as multimoding. Multimoding is illustrated in Fig. 13 which shows the major components of the waveguide system and some of the possible points of mode conversion. The delay ripples are a result of the unequal transit times for different modes of transmission in the circular waveguide. The spurious mode may be excited by incorrect antenna orientation, irregularities in the feed horn, excessive deformation of the flexible waveguide, and misaligned waveguide joints. Reconversion to the dominant mode may occur at the polarizing networks or, after reflection, back at the top of the run in one of the elements previously mentioned.

6.15 Tables B and C list conversion factors which relate the ripple spacing in megacycles to waveguide length in feet in WC281 circular waveguide. Table B is for the case of two reflections and the conversion numbers are given for all possible modes of transmission. Table C is for the case of multimoding as illustrated in Fig. 13. It is assumed in this table that the dominant mode is TE 11 and the conversion factors are for each of the other modes interacting with the TE 11 mode.

TABLE B — MULTIPLE REFLECTIONS IN WC281 WAVEGUIDE

MODE	GIGACYCLES			
	10.7	11.7	5.925	6.425
	CONVERSION FACTOR			
TE 11 (Dominant)	479	481	447	454
TM 01	469	473	413	426
TE 21	455	461	357	380
TL 11, TE 01	432	442	248	297
TE 31	419	431	158	239
TM 21	377	398		
TE 41	368	391		
TE 12	367	390		
TM 02	356	382		
TM 31	297	337		
TE 51	294	335		
TE 22	269	316		
TM 12, TE 02	237	294		
TE 61	172	254		
TM 41	157	245		
TE 32		198		
TM 22		136		
TE 13		110		
TE 71		99		
TM 03		76		

Example: TE 11 mode at 11.7 gigacycles, 4-mc spacing between ripples implies:

$$\frac{481}{4} = 120.2 \text{ feet between reflections.}$$

TABLE C — MULTIMODING IN WC281 WAVEGUIDE				
MODE	GIGACYCLES			
	10.7	11.7	5.925	6.425
	CONVERSION FACTOR			
TM 01	47,160	57,460	10,810	13,610
TE 21	18,140	22,290	3,510	4,640
TM 11, TE 01	8,840	11,020	1,110	1,720
TE 31	6,690	8,420	486	1,010
TM 21	3,570	4,650		
TE 41	3,180	4,170		
TE 12	3,150	4,140		
TM 02	2,790	3,710		
TM 31	1,570	2,250		
TE 51	1,530	2,200		
TE 22	1,230	1,850		
TM 12, TE 02	941	1,520		
TE 61	539	1,080		
TM 41	469	1,000		
TE 32		674		
TM 22		378		
TE 13		285		
TE 71		250		
TM 03		179		

Example: TE 61 mode at 10.7 gigacycles, 6-mc spacing between ripples implies $\frac{539}{6} = 89.8$ feet of waveguide between points of conversion.

RIPPLE AMPLITUDE VERSUS RATIO OF DESIRED AND UNDESIRED SIGNALS

6.16 If the delay pattern shows a nearly sinusoidal ripple as a result of one of the mechanisms given in the previous parts, an estimate may be made on the level of the delayed or undesired signal by means of the following formula.

Loss to undesired component

$$= 66 - 20 \log \tau_{max} S$$

Where τ_{max} is the peak-to-peak amplitude of the delay ripple measured in nanoseconds and S is the ripple spacing in megacycles.

Example: A ripple is found with a peak-to-peak amplitude 5.0 nanoseconds and a spacing of 4 mc

$$66 - 20 \log(20) = 40 \text{ db.}$$

In this case, the delayed signal level is 40 db below the desired signal.

6.17 All of the delay ripple mechanisms listed in this part may arise at either the transmitting or receiving end of a radio hop. Sometimes it will be possible to locate the site of the difficulty quickly on the basis of ripple spacing. This is true where the two installations use widely different lengths of waveguide. In other cases, it may be necessary to use attenuators to break up an echo path on one end of a hop while work on the other end of the hop continues.

7. RETURN LOSS TESTING

7.01 In cases where noise loading tests show a flat bottom in the top channel noise loading curve and where no delay measuring set is available, it may be possible to locate reflections in the waveguide system by means of return loss measurements. These tests may prove valuable in locating defective waveguide components or foreign objects in the waveguide run but will be of little or no use in cases of multimoding or loosely-coupled stray resonances (see Fig. 11 and 12).

7.02 Techniques for the return loss testing of the waveguide runs are included in Section 402-400-510. In each case, the preferred technique is to build up the waveguide run a section or two at a time in each case replacing the following waveguide with a known good termination. It must be noted that the transmitting antenna return loss may be measured properly only at the transmitter operating frequency and the receiving antenna may not be measured at all if radiation on unauthorized frequencies is to be avoided. The waveguide runs themselves may be measured at any frequency provided that the far ends are capped off with terminations.

7.03 Table D lists the allowable and the expected return loss values for the various waveguide components. The actual return loss measured at the end of a waveguide run will be the vector resultant of the various reflections within the run. Taking a simple example of two reflections each with a return loss of 24 db and assuming no loss in the waveguide run, the measured return loss will vary from 18 db to greater than 40 db as a function of frequency. The ripple rate will correspond to the rates found in Table A and will be a direct function of the physical spacing between reflections.

7.04 In an example installation, the return loss values might build up in the following way:

(a) The return loss of 3 feet of indoor waveguide and a pressure window is found to have a return loss of 35 db at 11.3 mc which does not vary appreciably with frequency changes of ±10 mc.

Discussion: The return loss from a single, wideband reflection does not depend upon waveguide position except for the transmission loss of the waveguide itself. The standing waves seen by a slotted line are not "seen" by a return loss set.

(b) The above components are again measured but in addition, 30 feet of WR90 waveguide and a flexible bend are added. The

TABLE D — WAVEGUIDE COMPONENT TYPICAL RETURN LOSSES AND TRANSMISSION LOSSES

COMPONENT	RETURN LOSS LIMIT (db)	RETURN LOSS TYPICAL (db)	TRANSMISSION LOSS ONE WAY (db/ft)
WR90 Waveguide 11-gigacycle Band			
Rigid Waveguide	40/flange	50/flange	0.04
Flexible Waveguide	31/section	35/section	0.07
90° Bend		35	
Pressure Window	32.5	35	
KS-19529, L30 Antenna	23	27	
WR159 Waveguide 6-gigacycle Band			
Rigid Waveguide	40/flange	50/flange	0.014
Flexible Waveguide		35/section	0.044
90° Bend	42		
Pressure Window	35		
1330A Filter	31		
KS-19529, L30 Antenna	23	27	

return loss is found to have ripples at a spacing of 13 mc with worst values of 31 db.

Discussion: The return loss from two wideband reflections separated by some distance in a waveguide will ripple between two values which result from the addition and cancellation of the reflected waves. The phase relationship between the two reflections will be a function of frequency and physical spacing. In this case, Table A may be used to determine the physical spacing as follows:

WR90, 11.3 gigacycles, TE 10 Mode, # = 401

$$\frac{401}{13} = 30.8\text{-ft spacing.}$$

The magnitude of the second reflection may be estimated by the following procedure. Note the return loss of the first reflection (35 db). Note the value of the peaks from two reflections (worst value) (31 db). Take the difference (4 db) and enter the conversion chart as shown in Fig. 14 by the line marked Step 2. Read the return loss correction +4.5 from the vertical scale and add to the previous value of 35. In this example, the degradation of 4 db corresponds to a new component of return loss which is 4.5 db greater than the previous worst value. The 35 + 4.5 = 39.5 db is measured at the waveguide input. The actual return loss of the flexible section is somewhat worse than this value. To get the correct value, subtract twice the waveguide loss between the return loss test set and the reflection

$$39.5 - 66 \times 0.04 = 36.9 \text{ db.}$$

(c) The previous components are again measured and in addition, 60 feet of rigid waveguide and a KS-19530 antenna are connected. The worst return loss peaks now measure 20 db although many others are noted of less importance. The ripples are no longer uniformly spaced.

Discussion: The return loss from three or more wideband reflections at varying distances in a waveguide will ripple between high and low values depending upon the phase relations between the reflected signals. It is usually very difficult to pick out the separation distances when more than two reflections are involved. It is possible, however, to estimate the value of the return loss of the added components as demonstrated previously. In this case:

31 − 20 = 11 − degradation

from Fig. 14, Step 3, new component −8 db

31 − 8 = 23 db

and allowing for the waveguide losses,

$$23 - (186 \times 0.04) = 15.6 \text{ db.}$$

This result indicates a deformed waveguide section, a foreign object, or some other damage beyond the point measured in Step 2.

This example illustrates a way in which return loss testing may be used to find defects in waveguide runs. It must be stressed that numerous antenna system faults which affect noise loading performance because of delay distortion cannot be identified by return loss tests.

8. CONCLUSION

8.01 This practice differs from most Bell System Practices that discuss measurements and analysis made on a transmission medium. An effort has been made to avoid reducing this practice to a set of testing steps and limits to be imposed on the corresponding measurements. A short haul microwave radio system is a complicated collection of equipment which under normal circumstances can be maintained and operated with a minimum of complicated test equipment. When situations occur, either during installation or while the system is being used for service (such complicated troubles seldom occur suddenly after service is established) where normal maintenance and trouble shooting techniques are inadequate, the more sophisticated methods discussed herein are required. To repeat an earlier statement in this section: the purpose of this and other heavily loaded system practices is to prescribe tests and adjustments which will permit optimum performance from each radio hop within the capability of the equipment. This does not mean that 25 dbrnc0 or better will be realized on each hop, but when such performance is not obtained, the tests will indicate the reason.

9. DERIVATIONS OF FORMULAS

9.01 This part contains the derivations of several formulas used in the text of this practice. They are included here as reference material for the engineer who may want to fit them to other systems, test equipment, or conditions of loading.

TRANSMISSION LEVEL

9.02 For 600-circuit loading, and assuming no pre-emphasis, the transmission levels shown in Fig. 14 are obtained with TL-2 or TM-1.

TALKER POWER

9.03 The approximate average power in an active telephone channel is taken to be −10 dbm0. Allowing for an activity factor of 25 percent, the average power is approximately −16 dbm0.

NOISE LOADING, POWER DENSITY, AND TOTAL POWER

9.04 The power level of −16 dbm0 per 4-kc wide channel is taken to be −17 dbm0 per 3 kc. In terms of annoyance, the message simulating signal level equals

−17 dbm0/3 kc + 88 = 71 dbrnc0 or
−17 dbm0/3 kc + 82 = 65 dba0.

The total power used to simulate 600 talkers is:

−16 dbm0 + 10 log 600 = +11.8 dbm0 total

and at the TL-2 or TM-1 transmitter input this power =

+11.8 dbm0 −35.5 (TLP) = 23.7 dbm.

The reference drive is then taken to be −24 dbm, a level which produces a peak frequency deviation of approximately 5 mc.

CONVERSION OF IDLE NOISE MEASURED AT RECEIVER OUTPUT TO TELEPHONE CIRCUIT NOISE

9.05 Noise measured by the 37B TMS:

+15.0 (conversion to 0 TLP)

$+10 \log \frac{3000}{500}$ (correction for bandwidth)

+1 (detector characteristic)

+88 (conversion to dbrnc0)

= Circuit noise in dbrnc0

or 37B TMS reading + 112 db = circuit noise in dbrnc0

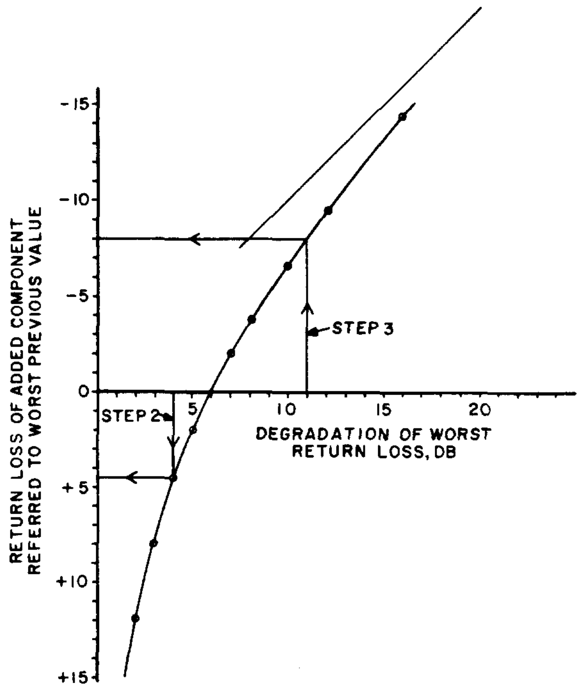

Fig. 14 – Conversion Chart for Return Loss Ripple Amplitude

NOISE AT RECEIVER OUTPUT DUE TO NOISE FIGURE

9.06 Deviation of carrier due to noise falling in a 1-cycle bandwidth at the receiver input =

20 log (rms frequency deviation) =

−174 + NF − (carrier power)

−3 db + 20 log (baseband frequency).

Note: The 3-db factor converts the carrier power to peak carrier power.

The receiver gain is such that an rms deviation of 4 mc yields an output of +9.5 dbm. Noise in 1 cycle at receiver baseband output:

$$= 9.5 \text{ dbm} + 20 \log \frac{\Delta f \text{ rms}}{4 \text{ mc}}$$

$$\text{NF } \underbrace{+9.5 - 174 - 3} - (\text{received carrier}) \quad +20 \log \left(\frac{\text{baseband frequency}}{4 \text{ mc}} \right)$$

and when measured with the 37B TMS,

$$= -167.5 - (\text{receiver carrier}) + 20 \log \frac{\text{baseband frequency}}{4 \text{ mc}}$$

$$+3 + 10 \log 500 - 1.$$

Where 3 db accounts for two sidebands producing noise at the same frequency, 10 log 500 converts to 500-cycle bandwidth and 1 db is allowed for the detector characteristic.

$$\left(\begin{array}{l} \text{Noise at the receiver} \\ \text{baseband output due} \\ \text{to noise figure and} \\ \text{measured by the} \\ \text{37B TMS} \end{array} \right) = -138.5 + \text{NF} - (\text{received carrier}) + 20 \log \left[\frac{\text{baseband frequency}}{4 \text{ mc}} \right]$$