Tektronix

Digital Modulation Technical Series

ACP Measurements on Amplifiers Designed for Digital Cellular and PCS Systems

The digital cellular and PCS industry places high demands on power amplifiers. Whereas in the past, two-tone third-order intermodulation testing may have been used to assess amplifier non-linear effects, there is an increasing case for using the genuine digitally modulated signal as a stimulus. This obviates the need to try to relate the systems specification parameters, such as Adjacent Channel Power, to other measures such as two-tone testing.

This technical brief examines the issues surrounding the generation of suitable signals for performing adjacent channel power measurements on components for use in IS-95 (CDMA) systems as well as methods of measurement.

I. Introduction

In connection with IS-95, there has been much use of band-limited noise signals to provide a spread-spectrum-like signal for use as a stimulus for ACP measurements. Measurements of the probabilities of exceeding a particular peak-to-average ratio can vary by a factor of 100:1 between noise (AWGN) and a Walsh-coded CDMA channel.

In addition, experiments have shown that the peak-to-average ratio of Walsh-coded signals can depend not only on the number of codes used, but also on the actual choice of codes. This, in turn, leads to variations in the ACP results due to the dependency of peak amplitude on the code selection used.

From a measurement perspective, absolute power measurements have shown that the errors attributable to widely used sample detection techniques vary for different combinations of Walsh codes. Results gained by use of an RMS detector will be shown for both noise and Walsh coded signals. Additionally, it will be shown that for relative measurements, the use of an RMS detector can lead to faster, more repeatable, and more stable measurements than could be achieved using sample detection techniques.

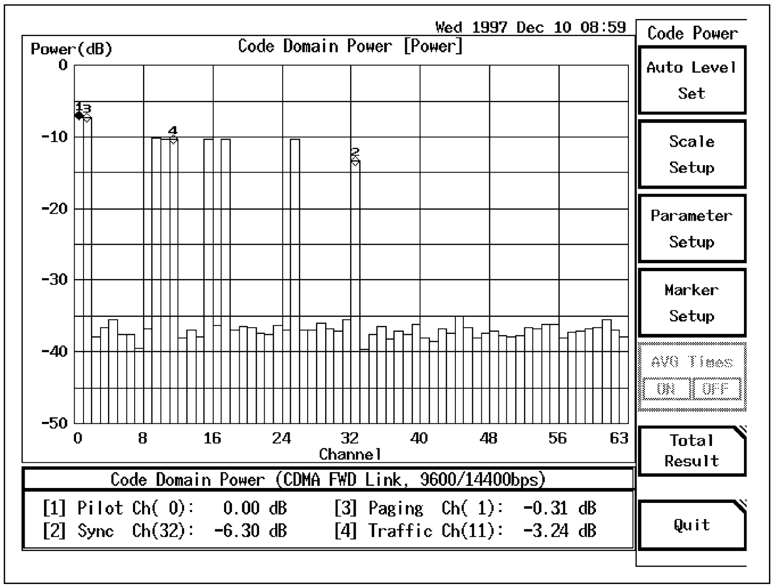

Figure 1. Code-domain power – base station. (Marker measurements relative to pilot.)

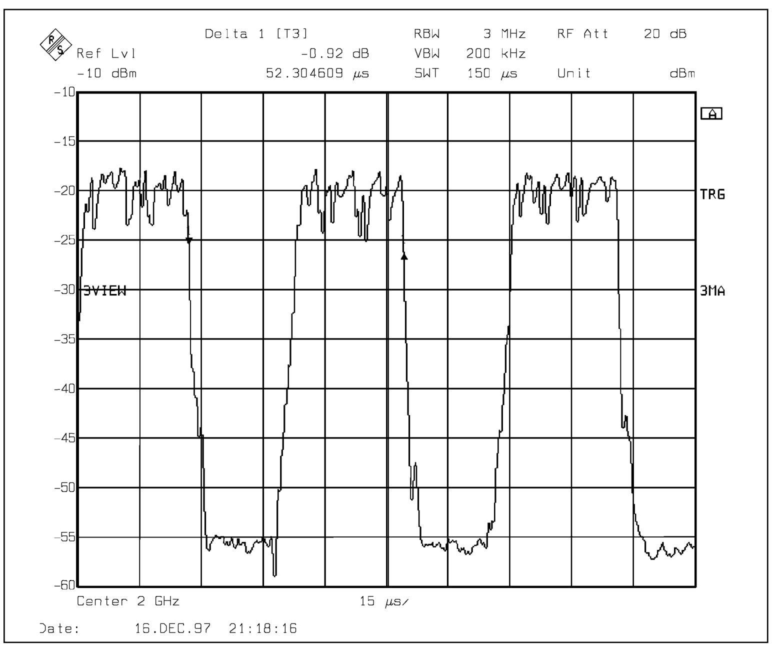

Figure 2. Additive superposition of Walsh-coded data, pilot, and sync (short code spreading off and data = 0).

II. Generation of Signals

The IS-97 forward-link signal is a summation of all traffic channels plus pilot, sync, and paging channels. To represent a lightly loaded forward link, IS-97 defines the waveform in Table 1. The model requires pilot, sync, paging, and six traffic channels.

Channel	Code Channel	Power, Linear	Power, Log/dB	Comments
Pilot	00	0.2000	–7.0	
Sync	32	0.0471	–13.3	1/8 rate
Paging	01	0.1882	–7.3	full rate
Traffic	Variable code channels (8 to 31 and 33 to 43	0.0941	–10.3	full rate
Traffic		0.0941	–10.3	
Traffic		0.0941	–10.3	
Traffic		0.0941	–10.3	
Traffic		0.0941	–10.3	
Traffic		0.0941	–10.3	

Table 1. IS-97 Base Station Model.

A graphical representation of this model is seen in Figure 1, where code-domain power is represented. However, the Walsh codes for traffic channels in the model are defined as variable. Careful selection of Walsh codes can create widely different RF envelopes, due to the additive super-position of Walsh-coded information in the RF modulation process. To illustrate this, a special case is demonstrated in Figure 2, where Walsh codes 0 and 32 are transmitting zeros simultaneously, with short code off.

Walsh codes 0 and 32 are similar; e.g., Walsh 0 is 64 zeros, and Walsh 32 is 32 zeros followed by 32 ones, so that an RF envelope in the shape of a square wave is created. The period of the waveform is the chip rate x 64 or 52 mS. As additional data channels are summed, this waveform becomes increasingly random. However, due to similarities in Walsh codes, it doesn't produce a Gaussian distribution of amplitudes.

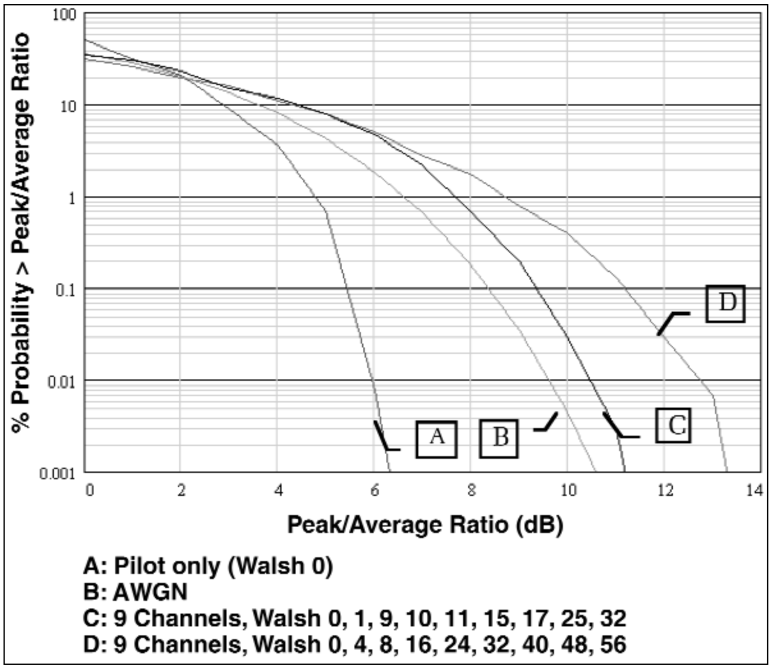

Figure 3. Cumulative distribution functions.

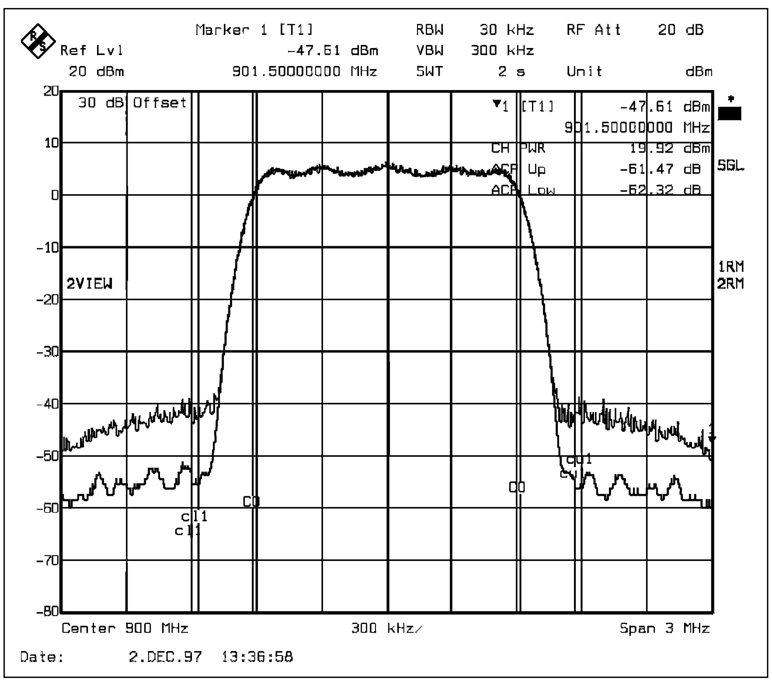

Figure 4. ACPR comparison, CDMA signals. a) Pilot only; b) nine-channel signal, Walsh codes 0, 32, 4, 8, 16, 24, 40, 48, and 56.

Examination of the Walsh codes in IS-95 finds that Walsh codes 0, 4, 8, 16, 24, 32, 40, 48, and 56 have greatest similarity. Use of these channels will create RF envelopes with higher probabilities of large peak/average ratios.

The data presented in Figure 3 demonstrates the differences in peak-to-average ratio that can result from selection of particular Walsh codes in a forward link signal as described by Tiepermann (Reference 1). In the case of a pilot signal only (Walsh 0), a simple QPSK waveform is produced, resulting in a peak/average ratio of 6.4 dB, due to the overshoot effects produced by the required IS-95 filter.

Both of the nine-channel waveforms in Figure 3 (c and d) are valid representations of a lightly-loaded forward link, but are very different in the amount of stress applied to an amplifier.

It’s also seen in Figure 3 that Walsh-coded waveforms are different from AWGN. Comparing the peak/average ratios, we see that the peak/average ratio of AWGN, at a probability of 0.001%, is 10.5 dB, and that of the two nine-channel CDMA waveforms are 11.1 and 13.6 dB. To approach the peak levels of the CDMA waveforms at similar probabilities, it’s necessary to increase the average power of the AWGN waveform by 0.6 dB and 3.1 dB, respectively. Increasing the average level in order to create similar peaks leads to an overdrive condition to the device under test.

Based on the information above, a test was performed to determine the effects of two different CDMA base station signals on power amplifier Adjacent Channel Power Ratio (ACPR). A pilot-only versus nine-channel waveform with similar Walsh codes were used to illustrate differences that can occur due to stimulus technique. The amplifier under test was rated at 35 dBm with 1 dB compression point, and output power was established at 20 dBm channel power.

In Figure 4, at 885 kHz offset, 30 kHz measurement bandwidth, an ACPR difference of 13 dB (62 dBc versus 49 dBc) is seen due to

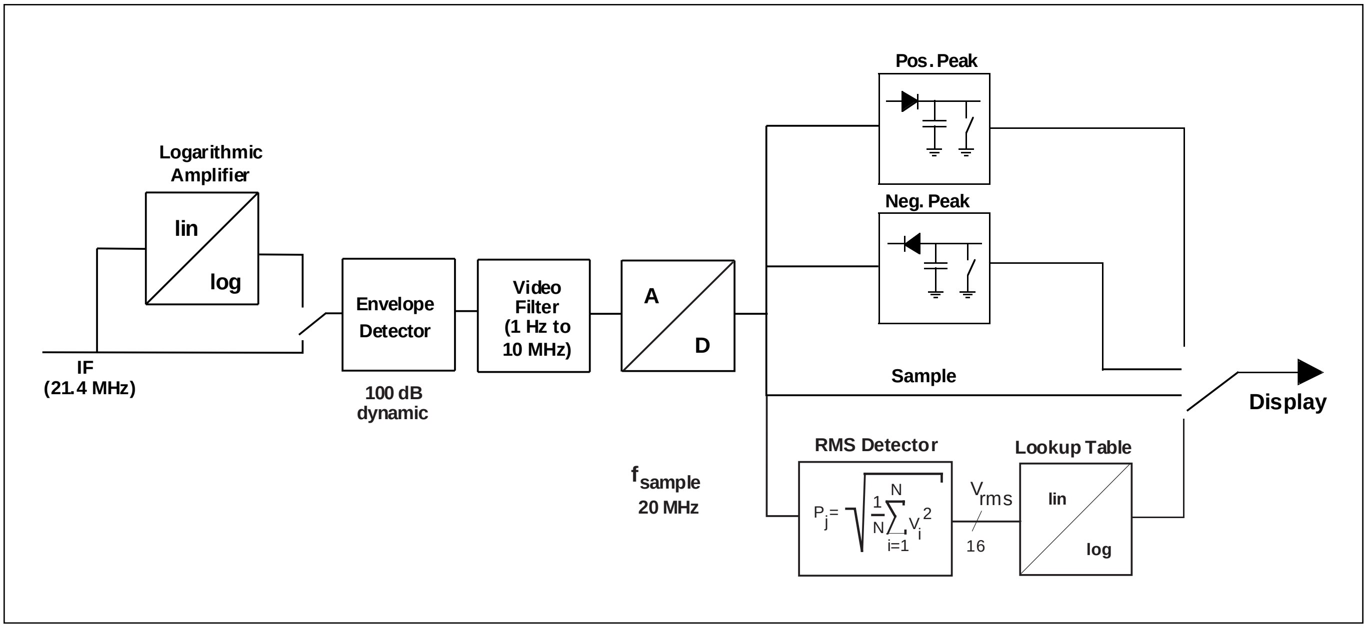

Figure 5. FSE spectrum analyzer block diagram.

the difference in peak/average value of the stimulus to the DUT.

III. Power Measurement

Prior to looking at the measurement of either power or adjacent channel power, it's necessary to explore how a spectrum analyzer makes power measurements.

Spectrum analyzers, as shown in Figure 5, use a logarithmic amplifier to attain sufficient dynamic range. This compresses the dynamic range of the signal. Following the logarithmic amplifier, a peak envelope detector removes the IF and delivers the envelope voltage of the IF voltage.

The video filter limits the bandwidth of the video voltage to reduce noise. It's equivalent to averaging the signal. However, due to the logarithmic scaling of the video voltage, averaging of the signal introduces amplitude errors for non-CW signals. When measuring white noise, the averaging causes the results to be 1.45 dB low. There is an additional 1.05 dB discrepancy derived from the difference between the average value and the power of white noise. This correction factor of 2.5 dB for noise measurement is well known and has been reported by Engelson (Reference 2). The same value applies for noise signals irrespective of whether the averaging is achieved by use of narrow video filtering or averaging over a number of sweeps. With other signal shapes, such as digitally modulated signals, the error will be different due to the different amplitude distribution of the signal. For power measurement, the use of narrow video filters or averaging is therefore not appropriate.

The above discussion has assumed the use of a sample detector, where one measurement sample has been obtained for each screen display point as the spectrum analyzer sweeps from start to stop frequency. An alternate technique called RMS detection may be used which obviates the need for use of narrow video filtering or trace averaging of logarithmic values to attain acceptable degrees of result stability.

Figure 5 shows the block diagram of the Rohde and Schwarz FSE spectrum analyzers. In addition to the traditional detectors, the FSE contains a power detector – the RMS detector. The workings of the RMS detector have been more fully described in Microwaves and RF (Reference 3). Here we'll concentrate on the differences between sample and RMS detection.

The sample detector samples once per screen display point. The RMS detector takes many more samples between points and uses all of these samples to calculate the RMS result, which is then represented as a single pixel. The number of samples per pixel depends on the sweep time, the number of screen points, and the sample rate of the analog-to-digital converter used. In the case of the FSE, the number of screen points is 500 and the sample rate is 20 MHz. The number of samples used in the calculation of RMS power-per-point is described in equation 1.

$$N = 20 \times 10^6 \times \text{Sweep Time}/(500 - 1) \qquad (1)$$

For a sweep time of 500 ms, this equates to 20,000 samples per point. These measurements have been achieved without use of a logarithmic amplifier in the IF stages and therefore do not need correction as described previously. In addition, the large number of samples/point obviates the need for trace averaging or use of narrow video filtering to obtain stable results.

It’s important to estimate what the effective sampling rate is; that is, to determine what samples contribute to the power measurement result in an uncorrelated way. The choice of resolution filter determines the effect that the samples have on any RMS or average calculation. This can be estimated by looking at the auto-correlation function of the impulse response of gaussian type filters – the filter types typically used in spectrum analyzers. When the auto-correlation function value is low, the individual samples can be deemed to be contributing independent information to an RMS or average process. When the auto-correlation function value is at its maximum value of 1, the samples are not contributing independent information. For a sampling rate equal to the resolution bandwidth, the auto-correlation function has a value of approximately 0.03. At this level, all samples may be deemed to be uncorrelated. This would offer a worst case estimate for the effective sampling rate; equation (1) is modified as shown in equation (2) to give the worst case effective number of samples included in a power calculation.

$$N_{eff} = \text{RBW} \times \text{Sweep Time}/(500 - 1) \qquad (2)$$

Figure 6 shows the major differences between using sample and RMS detection methods. The top trace is a measurement of noise obtained using the FSE RMS detector with a sweep time of 2 seconds. For the FSE, this would mean 80,000 samples-per-point. Taking into account the resolution bandwidth of 300 kHz, the effective number of samples becomes 1,200. The bottom trace was produced using a sample detector with a 5 ms sweep time and 500 averages. In both cases, the resolution bandwidth was 300 kHz and the video bandwidth 3 MHz.

The observed stability of the sample detected trace is approximately ±0.5 dB and that of the RMS detected trace approximately ±0.1 dB. The result for the sample detected trace with 500 averages agrees very closely with predictions using Monte-Carlo simulation for white noise as shown in Figure 7. The practical result for the RMS detector is better than the ±0.2 dB predicted from the Monte-Carlo simulation for 95% confidence limits. This is not unreasonable, as the assumption for the effective sampling rate being equal to the resolution bandwidth was based on requiring an auto-correlation function value of close to zero; in practice, even with higher sampling rate/resolution bandwidth ratios, and hence higher auto-correlation function values, there

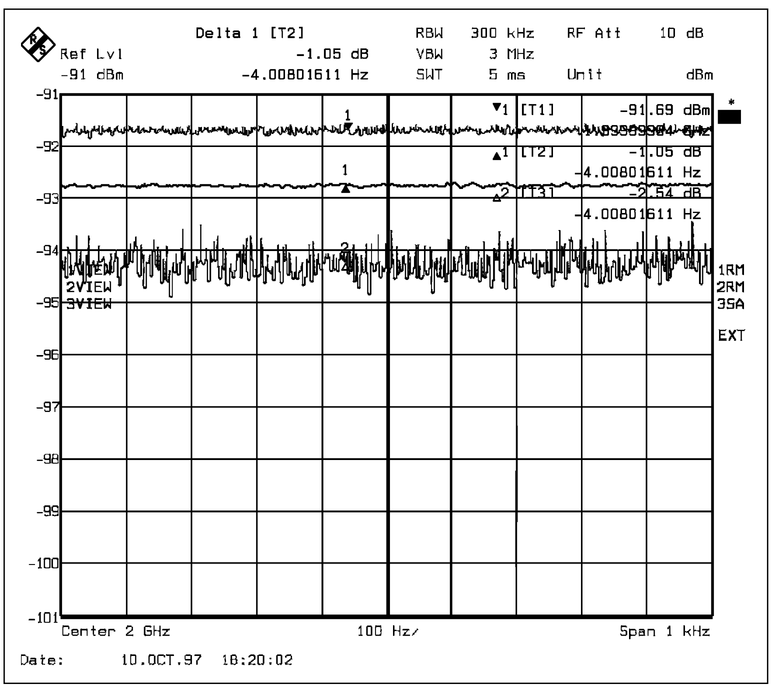

Figure 6. Comparing detectors.

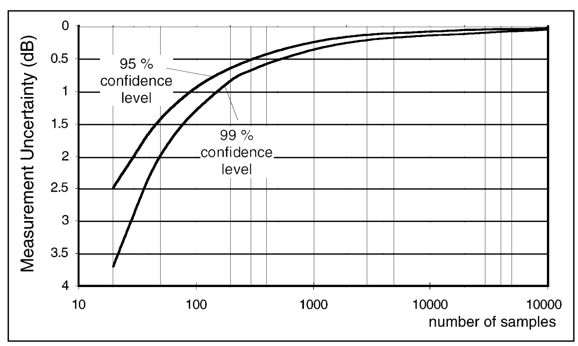

Figure 7. Theoretical sampling-dependent uncertainty (dB).

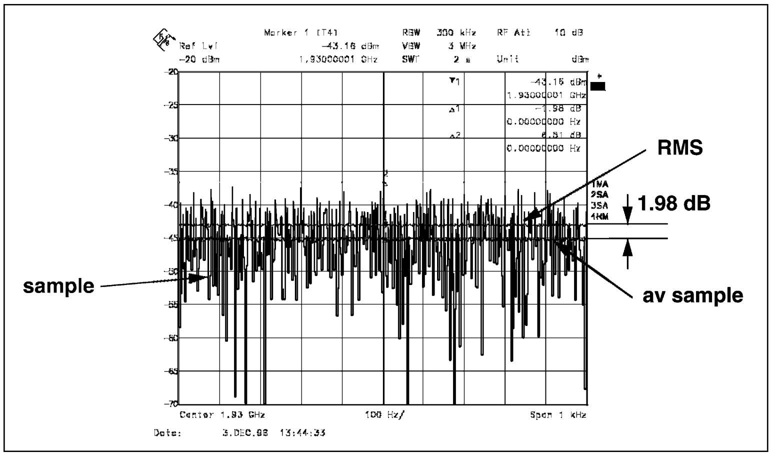

Figure 8. Differences for CDMA Pilot.

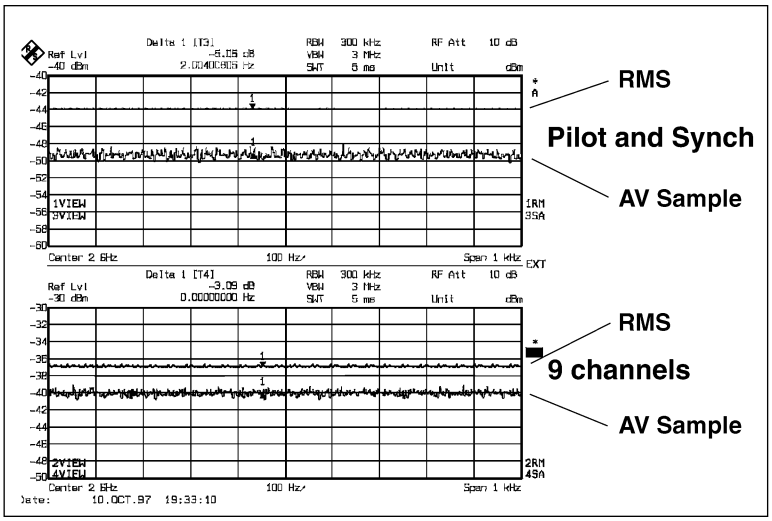

Figure 9. Changing amplitude distributions. For the pilot and sync case. The difference is approximately 5 dB; for the nine-channel case, the difference is 3 dB.

will be some additional contribution to the RMS or average calculations.

In addition to this improvement in stability, obtained in less time than for the average sampled result, the RMS-detected method requires no correction, whereas the sample-detected result is 2.5 dB low, exactly as predicted by theory. The middle trace shows the use of the RMS detector with narrow video bandwidth. This provides a result that is 1.05 dB lower than the power of the noise; this is due to the averaging effect of the video filter and illustrates that the average level of a white noise signal is 1.05 dB below the power of the signal.

The differences between sample and RMS detection changes depending on the nature of the signal to be measured. For a CDMA forward link pilot channel, the difference is 1.98 dB as shown in Figure 8.

When making measurements on CDMA signals with a mix of code channels, the differences between averaged-sample results and RMS detection again change due to the differing selections of Walsh codes. To illustrate, Figure 9 shows RMS and sample-detected results in the top screen area for pilot and sync channel and in the lower half for nine Walsh codes.

IV. Adjacent Channel Power Measurement

We have established the benefit of using an RMS detection method to accurately measure the power of a CDMA signal. This technique eliminates the need for Walsh Code related corrections and obtains stable results in a shorter period of time. We can now apply

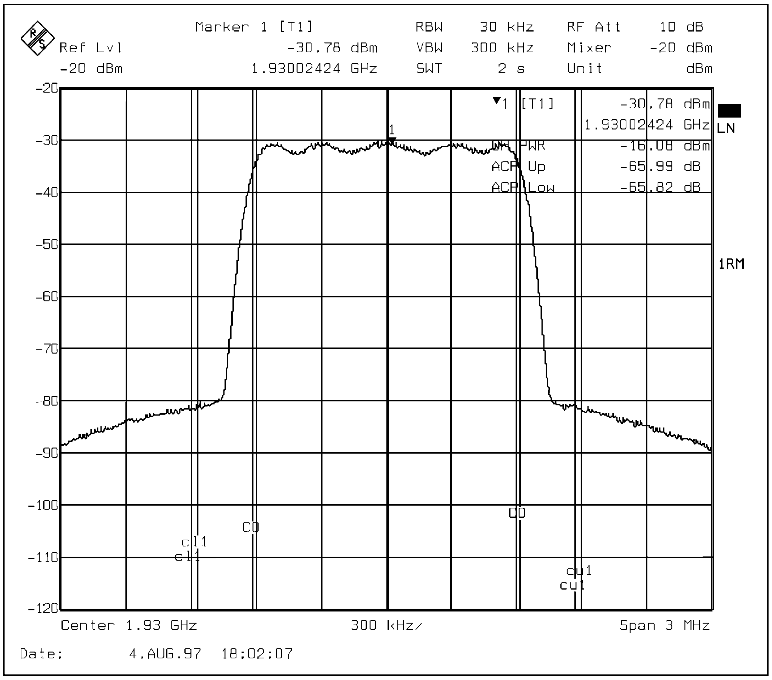

Figure 10. CDMA ACPR Measurement.

this technique to Adjacent Channel Power Ratio (ACPR) measurement.

ACPR is usually synthesized from two power measurements and, as such, the advantage that RMS detection provides in not needing to correct for averaging and signal amplitude distribution variation do not directly apply. The fixed correction would cancel out when making a ratio of the two results. This is not true in all situations as some ACP specifications rely on measurement of absolute power values. However, the ability of the RMS detection technique to obtain more stable results in a faster time is very relevant in all cases.

To determine ACPR, first the channel power of the CDMA transmitter signal is measured by integrating the power over a 1.23 MHz bandwidth. Then, the power at the required offset frequency is integrated over 30 kHz to measure the power in a particular adjacent channel. ACPR is the ratio of the two measurements. The term integration is used loosely in this case; the power measurements are more accurately described as the summation of the various power samples taken by the spectrum analyzer over the channel or adjacent channel bandwidths. Figure 10 shows the practical implementation of an ACPR measurement.

In the above example, there are 205 pixels used for the channel band and 5 pixels used for each of the adjacent channels.

Extending equation (2) to include factors for the number of pixels within the summation bandwidth and using the effective sample rate, the total number of samples included in the power measurement becomes:

$$N_{eff} = RBW \times Sweep\ Time \times (Ch\ BW)/Span \qquad (3)$$

Where: Ch BW = Channel bandwidth

Taking into account the resolution bandwidth of 30 kHz, this gives an effective number of 24,600 samples in the transmit channel and 600 samples in the adjacent channels. From the Monte-Carlo analysis (for white noise) it can be seen that the overall adjacent channel power ratio result will be dominated by the uncertainty associated with the 600 samples in the adjacent channel. From Figure 7, this can be seen to be approximately ±0.3 dB to 95% confidence limits. In practice, the variation from result to result has been better – in the range of ±0.1 dB, again indicating a higher effective sampling rate. This result was obtained with a sweep of only 2 seconds.

V. Conclusion

The measurement results presented have shown the need to be consistent in selection of measurement stimulus for ACPR measurement. The difference in amplifier ACPR between a pilot channel and a nine-channel case was up to 13 dB. Additionally it's concluded that the use of RMS detection, providing fast results independent of the amplitude distribution of the signal, is the way forward for absolute and relative power measurements on digitally modulated signals.

References

(1) Klaus Tiepermann, ***CDMA Signals, a Challenge for Power Amplifiers***, RF Design, July 1996.

(2) Morris Engelson, ***Modern Spectrum Analyzer Measurements***, JMS, Portland, OR, 1991, pp. 66-68.

(3) Josef Wolf and Bob Buxton, ***Measure Adjacent Channel Power With a Spectrum Analyzer***, Microwaves & RF, January 1997, pp. 55-60.

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