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MATRIX
TECHNICAL NOTES
MTN-110

EXTENDING THE LIMITS OF COMPOSITE INTERMODULATION DISTORTION MEASUREMENTS

1. INTRODUCTION

Intermodulation products that are the result of two frequencies are well known. The resultant second order and third order intermodulation products of two frequencies A and B are;

A+B, A-B, 2A, and 2B	For second order
2A-B, 2B-A, 3A and 3B	For third order

For many frequencies the predominant and important distortion products are;

A+B and A-B	For second order
A+/-B+/-C where A<B<C	For third order

When a system uses many equally spaced frequencies such as a Cable TV system, there may be hundreds or even thousands of distortion products. (REF. 1) These distortions are usually referred to as composite distortions because they are a composite of many discrete distortion products.

Just as in discrete distortions we have composite second order (CSO) for second order distortions, and composite third order, also known as composite triple beat (CTB) for third order distortions.

In a normal Cable TV frequency allocation the third order products fall in clusters around the carriers while the second order products are in clusters 1.25 MHz above and below the carriers.

These distortion products are usually measured by loading an amplifier with a given number of equally spaced CW carriers. One carrier is left off and the distortion products that exist in the band of the missing carrier are measured (REF. 2). FIG 1 is a block diagram of a typical distortion measurement configuration.

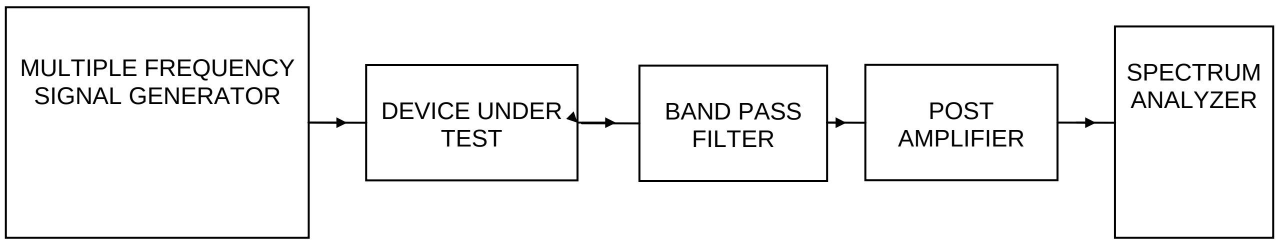


FIG. 1. TYPICAL CONFIGURATION FOR MAKING INTERMODULATION MEASUREMENTS

Because the frequencies of the carriers are not exactly equally spaced, the distortion products fall in a finite bandwidth. This mandates that a finite bandwidth be used for the measurements. The typical bandwidth used to make this measurement for a cable TV system is 30 KHz. (REF. 2) For other situations, such as a cellular telephone amplifier, the products may be much closer together and so allow the use of a narrower resolution bandwidth.

Given a fixed bandwidth and a fixed system noise figure, the magnitude of the smallest measurable distortion is also fixed.

It is convenient to define the minimum detectable distortion as that level of distortion equal to the noise level. This may not be a practical definition but simplifies some of the mathematics and conforms to other similar definitions.

Any measured distortion is the sum of the actual distortion and the system noise. If the distortion level is much greater than the system noise level, the measurement of distortion is straightforward. As the distortion level approaches the system noise level, the magnitude of the measurement is limited by the noise (See FIG. 2). The system noise is determined by the noise figure of the particular hardware and the configuration of the system. REF. 3 is an excellent tutorial source, and provides insight into some these results.

FIG. 2 shows the expected and measured third order distortion as a function of the actual distortion. Also shown are the thermal noise and intermodulation noise limits. Second order distortion would have a similar curve but with a different slope. The method presented here is directly applicable to second order measurements.

ACTUAL AND MEASURED DISTORTION AND NOISE

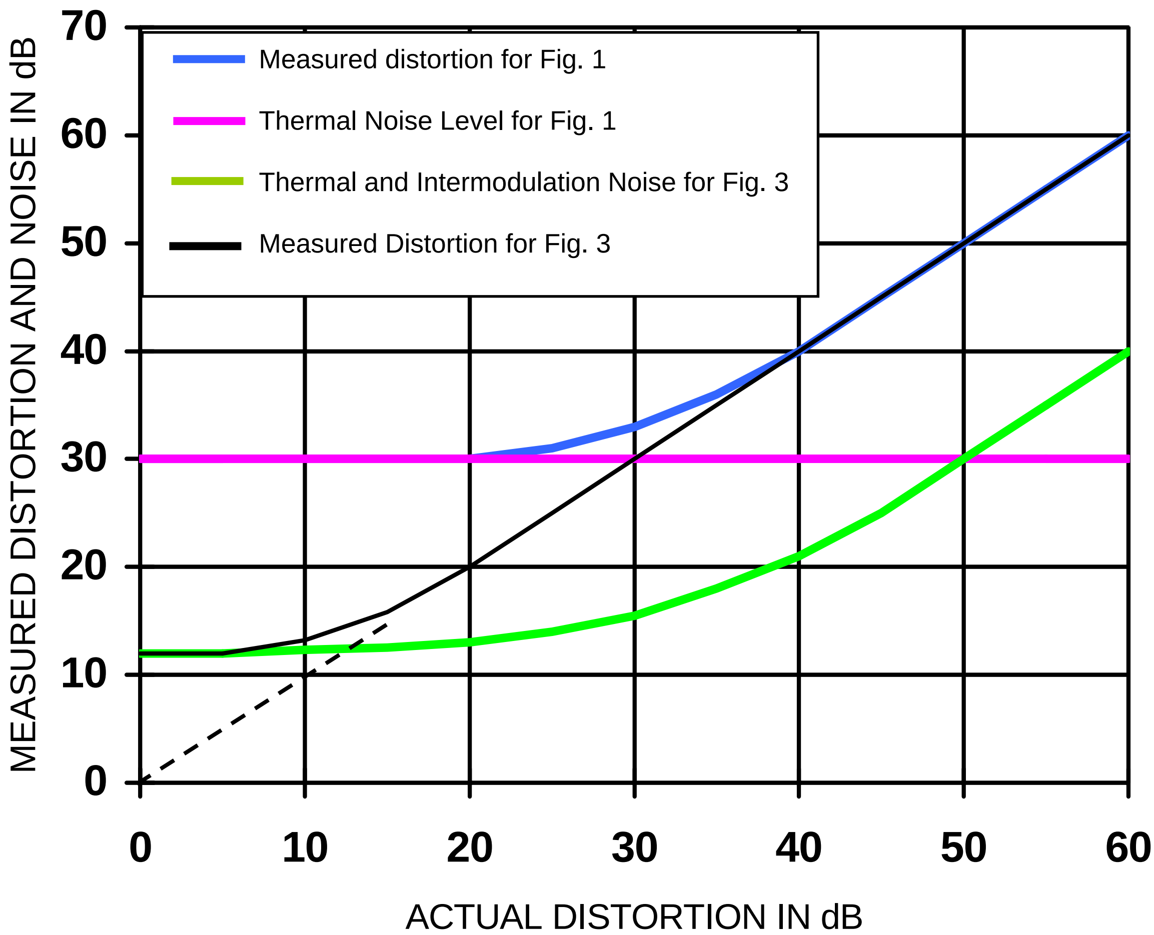


FIG. 2

It may seem unnecessary to be concerned with distortion levels that are near the noise levels. There are many instances where it is important to know the actual distortion magnitude even though it may be below the noise level. The most common example of this is in a cascade of amplifiers such as in a cable TV system. Here, as we add amplifiers to the cascade, the third order distortion products add as voltages while the noises add as powers. The net result is distortion levels, that are difficult to measure in a single amplifier, may be fatal in a cascade.

It would seem that a correction factor could be used to compensate for the noise addition to the distortion being measured. This was found to be impractical, first because low frequency fluctuations in both noise level and the system gain cause errors in the result, and second, as the distortion levels becomes smaller, the linear detector in the spectrum analyzer becomes a square law detector further degrading the quality of the measurement.

If the resolution bandwidth of the spectrum analyzer is reduced in an attempt to reduce the noise level and so improve the quality of the measurement, the number of distortion products being measured is also reduced, resulting in no improvement in the quality of the measurement.

Another method used to overcome the noise limitations is to make the measurements at a carrier level higher than normally used. For third order distortion, the distortion increases by 3 dB for every 1 dB increase in signal level. It should be possible to make a measurement at a higher carrier level, extrapolate back and calculate the distortion that would have existed at the lower carrier level. This method is successful with well-behaved amplifiers but fails if used, for example, with feed-forward amplifiers because the distortion products in feed-forward amplifiers are not well behaved.

2. OVERCOMING THE LIMITATIONS

The limitations in the measurement can be overcome if three changes are made.
(See FIG. 3)

1. Amplitude modulate the carriers with a square wave at some low frequency (1000 Hz).

2. Replace the linear detector (Demodulator) with a square law detector. (Also known as a quadratic detector and as a mean square circuit)

3. Follow the square law detector with a narrow band low frequency spectrum analyzer tuned to 1000 Hz.

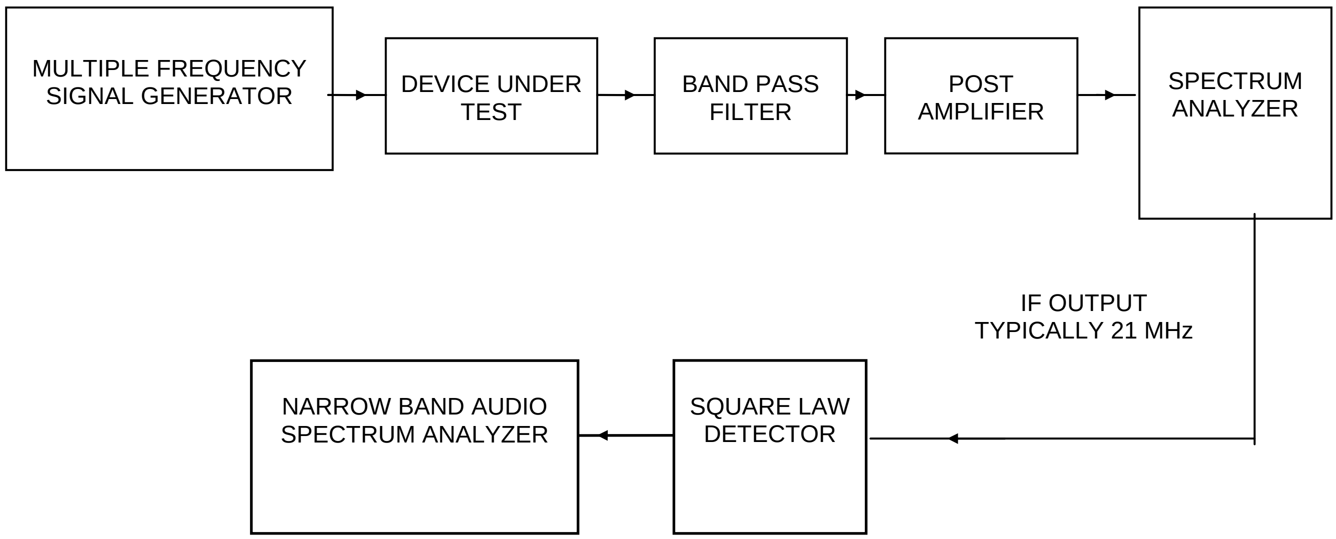


FIG. 3. THE MODULATION METHOD OF INTERMODULATION MEASUREMENTS

3. CIRCUIT DESCRIPTION

Modulation of the carriers is the key to this method. If all the carriers are 100% amplitude modulated with a 1 KHz square wave, the distortion products will also be 100% amplitude modulated with a 1 KHz square wave. If we extract the 1 KHz component of the demodulated signal with a narrow band device we see an improvement in the measurement, because reducing the bandwidth after the square law detector reduces the noise without altering the signal. The frequency of the modulation is chosen to be high enough so that low frequency variations are no longer a problem and low enough so that the signal we are trying to measure is not dispersed, by the modulation, to frequencies outside of the pre detection measurement bandwidth. (See FIG. 3)

At first glance it would seem that using a 1 Hz low pass filter at the output of the detector with no modulation would be equivalent to a 1 Hz bandwidth band pass filter at the output of the detector with modulation. If this were the case then the modulation method would show no improvement but rather a 3 dB degradation in performance. The main reason that the modulation method improves the measurement is related to the fact that noise produces a DC output from the second detector of the spectrum analyzer. It is the DC component of the detector output that limits the spectrum analyzer noise floor. Without modulation, small signals add only a small component to the DC of the second detector.

Reducing the video bandwidth or using video averaging does not improve the noise floor caused by the DC component. Furthermore the DC component contains low frequency fluctuations caused by noise and gain fluctuations. These low frequency fluctuations are commonly called 1/f terms because their magnitude increases as the frequency decreases.

The analysis of exactly how detectors behave with noise and noise like signals seems simple but in reality is very complex.

Fortunately there has been an analysis of a very similar problem. The circuit of FIG. 3 closely resembles part of the Dicke radiometer circuit of REF. 12.

The Dicke radiometer was designed to measure temperature by measuring the magnitude of the microwave thermal radiation from objects. Dicke's method was to alternately switch the receiver between an antenna and a termination. The output of the receiver was synchronously detected at the switching frequency.

The Dicke circuit is ideal for measuring small differences in noise levels. A modification of the Dicke receiver solves the noise floor problem.

The equations for the Dicke Radiometer indicate that the improvement in the measurement by using the circuit of FIG. 3 compared to a system using no modulation is;

$$\text{Improvement} = \frac{\sqrt{2}}{\pi}\sqrt{\frac{\text{(Pre Detection Bandwidth)}}{2\text{(Post Detection Bandwidth)}}} \qquad (1)$$

$$\text{Improvement(dB)} = 10\,Log\,\frac{\sqrt{2}}{\pi}\sqrt{\frac{\text{(Pre Detection Bandwidth)}}{2\text{(Post Detection Bandwidth)}}} \qquad (2)$$

If we use the radiometer equations with a pre detection bandwidth of 30 KHz, actually a noise bandwidth of 36 KHz, and a post detection bandwidth of 1 Hz we would expect to see an improvement of about 17.8 dB. This is very close to the actual measured values. See FIG. 2.

The example chosen uses a post detection analyzer bandwidth of 1 Hz. Other bandwidths as low as 0.1 Hz or as high as 100 Hz are quite practical. The measurement time is related inversely to the chosen bandwidth. With a 1 Hz bandwidth the measurement time is approximately 5 seconds.

4.1 SQUARE LAW DETECTOR

The reason for the choice of the square law detector is two fold. First, because the square law detector is square law throughout its operating range, it overcomes the small signal properties of the linear detector that becomes square law with signal levels near noise levels. (REF. 4) The second advantage of the square law detector is that its output voltage is proportional to the input power. This is a very important characteristic because the 1 KHz modulated distortions generate an output that is independent of the level of any noise also present. This is not true of the linear detector where the output is a function of the noise level. It is interesting to note that even if a perfect linear detector were available, the square law detector would be preferred for this application.

A further advantage if the square law detector is that because of the fact that **the square law detector responds only to power and is unaffected by high crest factors and is totally phase insensitive.** Crest factor is the ratio of peak to average ratio of the noise or distortion. Instruments such as the spectrum analyzer, that use linear detectors, are greatly affected by crest factor and as a result are one of the largest contributors to errors and discrepancies in the measurement of composite distortion.

FIG. 4 is a graph of Output voltage Vs Input power for two types of perfect detectors. Note that the square law detector has a linear power-voltage relationship. This fact simplifies the measurement method.

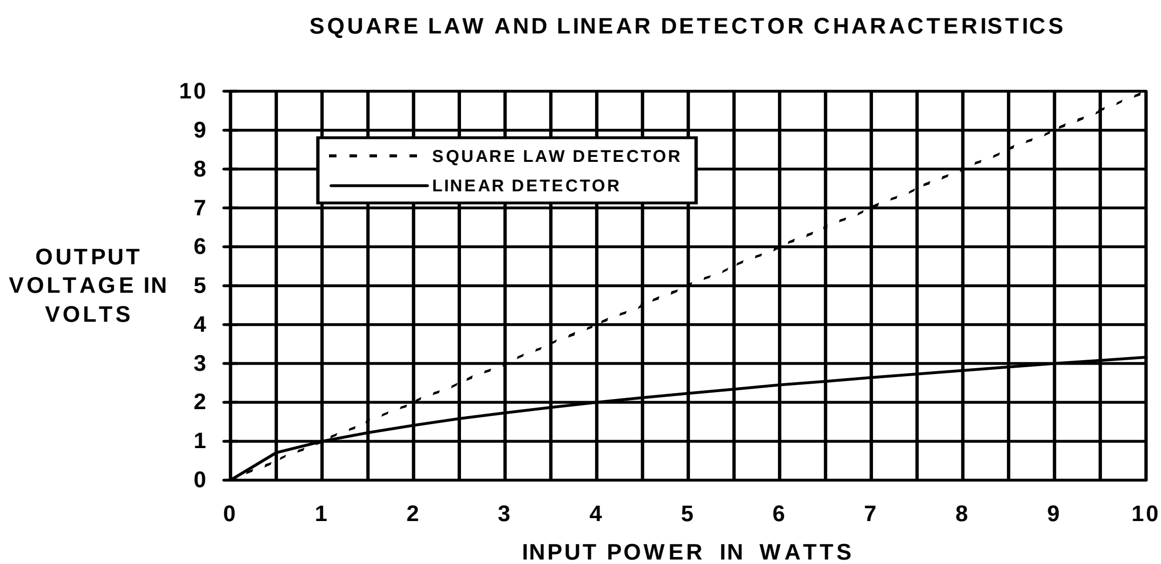


FIG. 4

Some further comments are warranted concerning the square law detector. We can write a simple expression for the square law detector.

$$V_o = (V_i)^2 \qquad (3)$$

V_o = Output voltage
V_i = Input voltage
P_i = Input power = $k_1(V_i)^2$
k_1 = Constant

and $$V_o = P_i / k_1 \qquad (4)$$

As expected the output voltage of the square law detector is linearly related to the input power. In a linear detector the output voltage is a square root function of input power. (See FIG. 4)

If the signals we are dealing with are modulated with a square wave of frequency F_m, we may write;

$$V_{om} = (V_i)^2 \,(k_2 \sin(2\pi F_m t) + k_3 \sin(6\pi F_m t) + k_4 \sin(10\pi F_m t) \ldots)$$

and $$V_{om} = (P_i) \,(k_2 \sin(2\pi F_m t) + k_3 \sin(6\pi F_m t) + k_4 \sin(10\pi F_m t) \ldots)$$

where $$(k_2 \sin(2\pi F_m t) + k_3 \sin(6\pi F_m t) + k_4 \sin(10\pi F_m t) \ldots)$$

is the Fourier series of a unit square wave.

and V_{om} = Output voltage with modulated input signal

The output audio analyzer is tuned to F_m so only the first term of the series is required. All of the constants in the analysis can be evaluated but it is not necessary to do so because we are dealing with ratios. (REF. 11)

We may now write;

$$V_{dm} = (P_{dm}/ k_1)(k_2)\sin(2\pi F_m t) \quad (5)$$

$$V_{dm} = (P_{dm})(k_5)\sin(2\pi F_m t)$$

where $k_5 = k_2 / k_1$

and V_{dm} = Voltage output at frequency F_m for modulated distortion input
P_{dm} = Peak of the modulated distortion power

In a similar way we may write;

$$V_{cm} = (P_{cm})(k_5)\sin(2\pi F_m t)$$

where V_{cm} = Voltage output at frequency F_m for modulated reference carrier input

and P_{cm} = peak input of the modulated carrier power

Note that the output voltage of the square law detector is a square wave with frequency F_m, 1 KHz in our case, and whose amplitude is proportional to the input power. The audio spectrum analyzer measures the magnitude of the fundamental frequency that is proportional to the magnitude of the square wave.

from above $V_{dm} = (P_{dm})(k_5)\sin(2\pi F_m t)$

and $V_{cm} = (P_{cm})(k_5)\sin(2\pi F_m t)$

but $P_d = k_6(P_{dm})$

and $P_c = k_6(P_{cm})$

where K_6 = constant

k_6 is a constant because both signals have the same waveform. We can combine k_5 and k_6 into k_7.

$$V_{dm} = (P_d)(k_7)\sin(2\pi F_m t)$$

and $V_{cm} = (P_c)(k_7)\sin(2\pi F_m t)$

from above $V_{dm}/V_{cm} = (P_d)(k_7)\sin(2\pi F_m t)/(P_c)(k_7)\sin(2\pi F_m t)$ (6)

or $V_{dm}/V_{cm} = P_d / P_c$ (7)

The audio analyzer can easily display;

$$20\log(V_{dm}/V_{cm})$$

and $$20\log(V_{dm}/V_{cm}) = 20\log(P_d/P_c) \quad (8)$$

We seek $$10\log(P_d/P_c) = \text{distortion to carrier power ratio in dB}$$

Therefore the distortion to carrier power is measured as;

$$P_d/P_c(dB) = 10\log(P_d/P_c) = (20\log(V_{dm}/V_{cm}))/2$$

and $$P_d/P_c(dB) = (20\log(V_{dm}/V_{cm}))/2 \quad (9)$$

This is the very nature of the square law detector. The choice of the square law detector has solved the problem of the linear detector but has forced a requirement of greater dynamic range on the 1 KHz measurement. To some extent the dynamic range at the output can be decreased by changing the gain at the input to the square law detector.

One minor point must be considered when measuring composite distortions. The standard method for composite distortion measurements uses a spectrum analyzer. The definition of carrier to composite distortion assumes that the measurement is made using a spectrum analyzer in the Log display mode. Spectrum analyzers however read noise and noise-like signals in error. Usually, when measuring noise with a spectrum analyzer a correction factor of about 2.5 dB is used. This correction factor of 2.5 dB is not used when measuring composite distortion. Because the square law detector is truly measuring the carrier power to composite distortion power ratio, the noise levels read with the modulation method must be decreased by 2.5 dB if agreement with the spectrum analyzer is desired.

> Composite distortion = $((20\log(V_{dm}/V_{cm}))/2) - 2.5$ dB
>
> (Note that $(20\log(V_{dm}/V_{cm})$ is read from the audio analyzer and has a negative value)

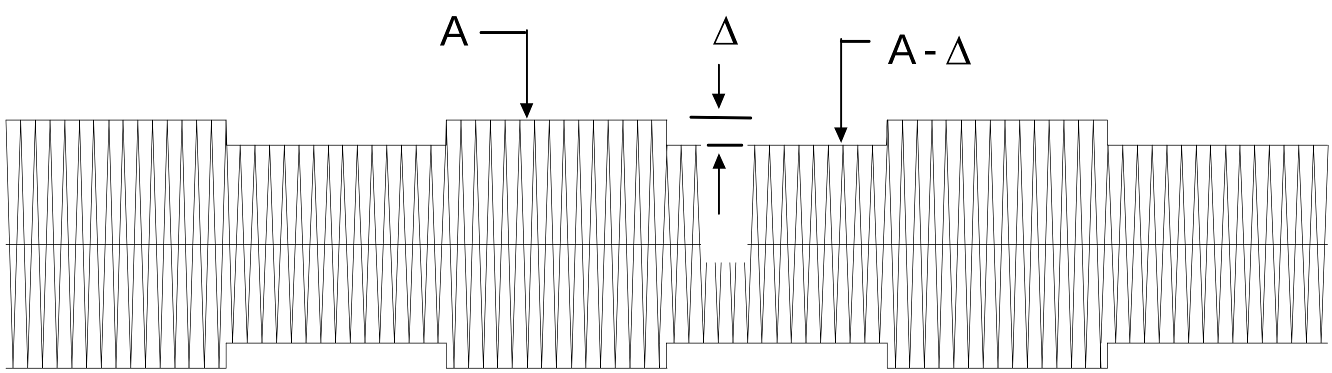
5. MEASUREMENT OF CROSSMODULATION

There are two definitions for cross modulation. One definition relates the modulation sideband magnitude to the CW carrier magnitude. The other more common definition relates magnitude of the modulation sideband to the sideband of a 100 % amplitude modulated carrier. The above definitions are further confused by the CATV definition of modulation being only downward modulation. For our example we will use the more common second definition of crossmodulation and the CATV definition of modulation.

The circuit shown in FIG. 3 was designed to solve the particular problem of composite distortion measurements. If we compare this circuit with the circuit used for crossmodulation measurements (REF. 13), we find the only difference is the type of detector used.

When measuring crossmodulation great care is taken to verify that the detector is operating in the linear mode. For composite distortion, great care is taken to operate in the detector in the square law mode. We will show that the square law detector can be used to make crossmodulation measurements with a technique similar to that of the linear detector except for a 6 dB offset.

FIG. 5 is a presentation of a carrier with some degree of crossmodulation.



A = CW carrier voltage
Δ = change in A due to distortion

FIG. 5

$$\text{Crossmodulation(dB)} \equiv 20\log \Delta/A \qquad \text{(See REF. 13)} \qquad (10)$$

The output of the square law detector with carrier level A input is;

$$k_5 A^2 \qquad (11)$$

With 100% square wave modulation, the output would be a square wave with magnitude;

$$k_5 A^2 \qquad (12)$$

With distortion present, the output of the square law detector is;

$$k_5 (A - \Delta)^2 = k_5 (A^2 + \Delta^2 - 2A\Delta)$$

$$\Delta^2 \approx 0 \qquad \text{for small } \Delta$$

$$k_5 (A - \Delta)^2 \approx k_5 (A^2 - 2A\Delta) \qquad (13)$$

The magnitude of the square wave, at the detector output, caused by the distortion is (12) - (13) or;

$$k_5 A^2 - k_5 (A^2 - 2A\Delta) = + k_5 2A\Delta \qquad (14)$$

If we divide the result from (14) by (12) we have;

$$k_5 2A\Delta / k_5 A^2$$

or

$$2\Delta/A \qquad (15)$$

The audio analyzer can display 20log of the ratio of the distortion square wave to carrier square wave at the detector output as;

Audio analyzer reading = $20\log(2\Delta/A)$

$$20\log(2\Delta/A) = 20\log(2) + 20\log(\Delta/A) \qquad (16)$$

$$20\log(2\Delta/A) = +6\text{dB} + 20\log(\Delta/A)$$

$$20\log(\Delta/A) = 20\log(2\Delta/A) - 6\text{dB}$$

(Note that $20\log(2\Delta/A)$ has a negative value)

Crossmodulation is defined as; Crossmodulation $\equiv 20\log(\Delta/A)$

Crossmodulation = $20\log(2\Delta/A) - 6$ dB

(Note that $20\log(2\Delta/A)$ is read from the audio analyzer and has a negative value)

6. PRACTICAL IMPLEMENTATION

All of the items in the block diagram of FIG. 3 except the square law detector are commercially available (REF. 5, 6, 7). Here the spectrum analyzer functions only as a frequency converter and so must have available a low frequency IF output port. The square law detector was not available so one was designed for this application using a high frequency four quadrant multiplier (REF. 8, 9). For our particular example, the IF frequency was 21 MHz, but, operation at other frequencies proved to be satisfactory.

There are several choices for the type of low frequency analyzer. The best choice seems to be the FFT type because this type provides the greatest versatility in the choice of bandwidth and the availability of averaging. FFT type analyzers several window functions but the equivalent noise bandwidths may not be made clear. They do however provide a noise density function which displays noise in a 1 Hz bandwidth. For our example we used the spectrum display mode to measure reference and distortion magnitudes and the noise density mode to measure noise levels. The window function chosen was “flattop” because it provided best accuracy when measuring the reference.

For composite distortions, calibration is done by inserting a carrier on the measurement channel, 100% amplitude modulated by a 1 KHz square wave, and recording the magnitude of the 1 KHz at the audio analyzer. The carrier is now removed, all other carriers are modulated, and the magnitude of the 1 KHz, which is from the distortion alone, is again measured. One half the difference in dB (a negative number) - 2.5 dB is the composite distortion. FIG 2 shows the results of actual measurements on a typical amplifier.

For Crossmodulation, calibration is done by inserting a carrier on the measurement channel, 100% amplitude modulated by a 15.750 KHz square wave, and recording the magnitude of the 15.750 KHz at the audio analyzer. All carriers are now modulated with 15.750 KHz except the test channel that is operated CW. The magnitude of the 15.750 KHz is again measured. The difference in the measurements (a negative number) - 6 dB is the crossmodulation.

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