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DISTORTIONS IN THE OUTPUT SIGNALS OF CONVENTIONAL SPECTRUM ANALYZERS *

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ABSTRACT

We show that the output signals of conventional spectrum analysers contain distortions which basically originate from the signal processing performed inside the analysers' frequency converters. Total elimination of these distortions through normal filtering techniques is difficult owing to the closeness of some of their frequencies to the corresponding frequencies of the required signals. Simple design adjustments that can minimize these distortions are suggested.

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1. INTRODUCTION

Contamination or distortion of information signals in analog electronic systems is done by one or more of the following categories of noise: man-made noise, erratic noise from natural phenomena, fluctuation noise and noise related to signal-processing. In practice, man-made noise may be eliminated either by removing the noise sources or by making them harmlessly directional. Erratic noise from natural phenomena may be minimized through appropriate electro-magnetic shielding. As far as many communication and other electronic systems are concerned, fluctuation noises (which include thermal and shot noises) are justifiably assumed to constitute random, ergodic and gaussian processes. On the basis of this assumption, remedial techniques are formulated and implemented (e.g. Davenport and Root, 1958; Motchenbacher and Fitchen, 1973; Robins, 1982).

Distortions due to signal-processing as well as their practical remedies depend on the particulars of the processing involved (e.g. Njau 1985; 1987). Where the spectra of such distortions do not intrude into the required bandwidths, then they can be eliminated through appropriate filtering. However, if the spectra of the distortions encroach closely into the required bandwidths, it will be impossible to selectively eliminate them through normal filtering without introducing further distortions into the required signals. In this paper, we show that the outputs of conventional spectrum analysers incorporate distortions which are a result of signal-processing. The frequencies of these distortions are inevitably within the required bandwidths. Our analysis shows that it is difficult if not impossible to completely eliminate these distortions through normal filtering methods because their frequencies are too close to the wanted frequencies. Nevertheless, we suggest a practical technique by which these distortions can be at least minimized.

2. FORMULATION AND DISCUSSION

A typical spectrum analyzer may be represented by the block diagram shown in Fig. 1. Its operation may be summarized as follows. The ramp generator repeatedly produces ramp voltage at fixed sweep time or period T_s. During each complete sweep, the ramp voltage makes the voltage-controlled oscillator (VCO) output a constant-amplitude sinusoid $g(t)$ whose frequency varies from a certain reference frequency ω_0 up to a maximum frequency $\omega_0 + \omega_M$. Signal $g(t)$ is continuously multiplied by the input signal $f(t)$, the result of which is passed through a narrowband filter at fixed frequency ω_0 and with a fixed bandwidth Δ. The spectrum of the output $s(t)$ of the filter is what appears on the oscilloscope screen after some envelope detection as shown in Fig. 1. Thus the spectrum of $f(t)$, which is fully displayed over the oscilloscope screen during each time-length T_s, has an overall frequency range $0 - \omega_M$.

Effectively, the whole spectrum of $f(t)$ is divided and analysed into N successive frequency (resolution) cells such that $N = \omega_M / 2\pi\Delta$. During a complete sweep, each cell is allocated time–length $T = T_s/N$. Let us arbitrarily consider that the n^{th} cell (in a given sweep) is analysed from $t = t_n$ to $t = t_n + T$ and that the corresponding output from the multiplier is denoted by $R_n(t)$. We assume that during the period $t = t_n$ to $t_n + T$, the output of the VCO is a sinusoid at amplitude A and frequency ω_n. Now if β and α are the d.c. bias plus d.c. signal levels along the VCO–multiplier line and along the $f(t)$–carrying input line of the multiplier, respectively, then $R_n(t)$ may be resolved into a series of spectral components as follows:

$$
\begin{aligned}
R_n(t) &= \{\alpha + f(t)\}\{\text{Portion of } (\beta + A\sin\omega_n t) \text{ sandwiched} \\
&\quad \text{between } t = t_n \text{ and } t = t_n + T \text{ as well as time} - \text{domain} \\
&\quad \text{distortions resultant from the implicit truncation or time} - \text{limiting process}\} \\
&= \frac{2}{\pi}\{\alpha + f(t)\}\{\beta + A\sin\omega_n t\}\{\sin\omega_T t + \sin 2\omega_T t + \\
&\quad \frac{1}{3}\sin 3\omega_T t + \frac{1}{5}\sin 5\omega_T t + \frac{1}{3}\sin 6\omega_T t + \frac{1}{7}\sin 7\omega_T t + \\
&\quad \frac{1}{9}\sin 9\omega_T t + \ldots\} \\
&= \{\alpha + f(t)\}\{\beta + A\sin\omega_n t\}D(t) \qquad (1)
\end{aligned}
$$

where

$$\omega_T = \frac{\pi}{2T}$$

and

$$
\begin{aligned}
D(t) = \frac{2}{\pi}\{&\sin\omega_T t + \sin 2\omega_T t + \frac{1}{3}\sin 3\omega_T t \\
&+ \frac{1}{5}\sin 5\omega_T t + \frac{1}{3}\sin 6\omega_T t + \frac{1}{7}\sin 7\omega_T t + \frac{1}{9}\sin 9\omega_T t + \ldots\} \ .
\end{aligned}
$$

The time–domain version $H(t)$ of the entire spectrum displayed over the oscilloscope screen during each complete sweep may be deduced from Eq.(1) as

$$
\begin{aligned}
H(t) = \frac{\alpha\beta D(t)}{(0 \le \text{ frequency } \le \omega_M)} &+ \frac{\beta f(t) D(t)}{(0 \le \text{ frequency } \le \omega_M)} \\
&+ \frac{D(t)[\alpha + f(t)]}{(0 \le \text{ frequency } \le \omega_M)} \sum_{n=1}^{N} A\sin\omega_n t \qquad (2)
\end{aligned}
$$

An analysis of Eq.(2) shows that the second and third sets of terms on the right–hand–side of this equation produce peaks at the expected frequencies except that, due to the influence of $D(t)$, these peaks are inevitably widened or spread out more than ideally required. This frequency spreading is clearly due to the spectral components of $D(t)$ that have largest amplitudes and least frequencies. The role played by the remaining components is simply to introduce some (usually) very small peaks which spread around

the peak of each expected or correct frequency, getting fainter as they distance away from the latter peak. The cluster of small peaks around each correct peak together with the frequency spreading mentioned above are distortions which stem from the time–limiting imposed upon the part of $g(t)$ corresponding to each frequency cell. As long as time–limiting is imposed upon $g(t)$, these distortions will inevitably be incorporated into the output of the narrowband filter. This is not surprising because time–limiting of a signal by itself introduces significant distortions (e.g. Lathi, 1965; Stark and Tuteur, 1979). Therefore, although at a given time during a sweep the narrowband filter may give out a sinusoid at the expected frequency, the very fact that this sinusoid is time–restricted to within a finite cell introduces the type of distortions just mentioned, and these distortions do accompany the filter output signal onto the oscilloscope screen. Of course, apart from frequency–widening each expected peak in the output spectrum, these distortions only manifest themselves as (often negligible) series of minute peaks on either side of each expected peak.

The main problem lies with the distortions caused by the first set of terms on the right–hand–side of Eq.(2). Firstly we note that $\alpha\beta D(t)$ has no definite or non–accidental spectral relationship with the input signal $f(t)$. Secondly, the energy in $\alpha\beta D(t)$ is, to a large extent, concentrated in the frequency range $\sim \frac{\pi}{2T}$ to $\sim \frac{3\pi}{2T}$. Thus if $f(t)$ has no component whose frequency coincides with part of the latter range, then the entire spectrum of $f(t)$ as displayed on the analyser screen will show the expected or correct peaks plus at least one significant distortion peak located between frequencies $\sim \frac{\pi}{2T}$ and $\sim \frac{3\pi}{2T}$, depending on the resolution employed. This is illustrated in Fig. 2 which shows the spectrum analyser display of a single sinusoid from a signal generator set at 390 kHz. The centre peak on the figure corresponds to the expected or actual sinusoid frequency while the relatively smaller peak at the low frequency end of the spectrum is a distortion peak located within the frequency range $\sim \frac{\pi}{2T}$ to $\sim \frac{3\pi}{2T}$ as calculated from the adjustment setting of the commercial spectrum analyser used. We operated the latter at a fairly low resolution in order to compress the closely packed distortion peaks within the frequency range $\sim \frac{\pi}{2T}$ to $\sim \frac{3\pi}{2T}$ into a single peak as shown on Fig. 2. Besides, the latter peak is given prominence in the figure because we intentionally set the amplitude of the sinusoid from the signal generator moderately low.

The distortions generated by $\alpha\beta D(t)$ in Eq.(2) may be minimized by making or adjusting the design of the frequency converter unit in the spectrum analyser in such a way that the VCO–generated signal is accompanied by as little d.c. component as possible on its way to the active device of the multiplier. With the currently popular brands of multipliers used in commercial spectrum analysers, it is difficult to completely rid the VCO–generated sinusoids of d.c. components. As a rule, successful operation of logarithmic multipliers upon a.c. signals requires that the latter be accompanied by d.c. components. This is inevitable since the inputs of these multipliers have to be unipolar. As regards variable transconductance multipliers, it is rather difficult to have them operate on analog signals without having d.c. components along their input channels. This is the case because quite commonly, a multiplier in this category has the configuration of an emitter-coupled differential amplifier in which one input signal controls the current along the emitter arm while the second input signal serves as a normal input signal for the amplifier.

3. CONCLUSION

We have developed and discussed details of distortions that are incorporated into the outputs of conventional spectrum analysers as a result of the signal-processing performed in the frequency converter units. The time-limiting processes employed in this signal-processing make it practically impossible to eliminate all the distortions. However, it is possible to minimize these distortions by making some adjustments in the design of the multiplier section of each frequency converter unit.

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Figure Captions

Fig. 1 Block diagram of a typical spectrum analyser.

Fig. 2 Spectrum analyser display of the spectrum of a sinusoidal signal at 390 kHz. The peak towards the low frequency end of the spectrum represents some distortions.

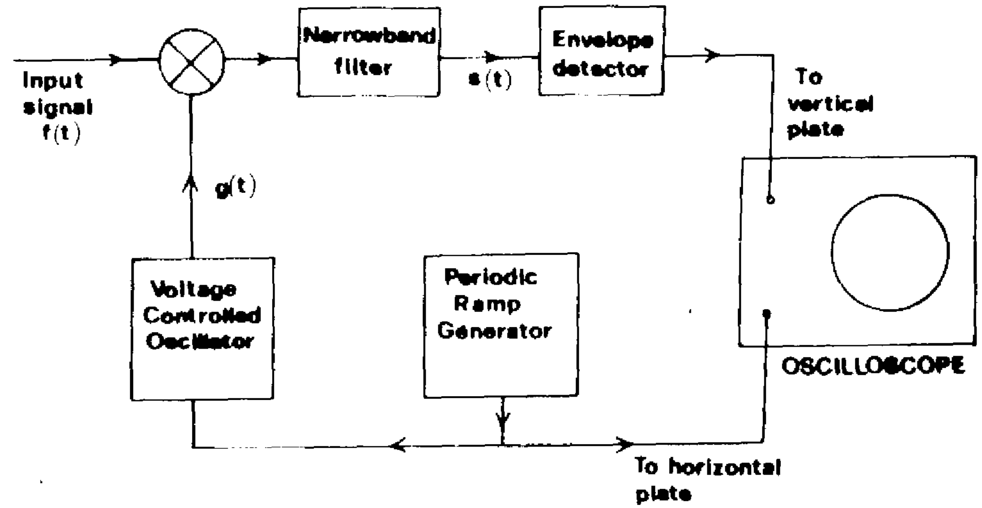

Fig. 1

Fig. 2

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