The LINC Transmitter

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A technique for producing linear amplification of a bandpass signal using both non-linear components and digital signal processing is the subject of this article. With this technique the bandpass signal is decomposed into two constant-amplitude phase modulated signals. These two phase modulated signals can be nonlinearily amplified and passively recombined to produce a linearly amplified signal.

In mobile radio systems, the relatively inefficient use of the spectrum by existing types of FM modulation like MSK, TFM, etc., has resulted in crowding on the available channels. They are still widely used because their constant envelope property is appropriate for using power efficient non-linear amplifiers. In the next generation of digital cellular radio systems, the use of Quadrature Amplitude Modulation (QAM) patterns will be required because they have higher spectrum efficiency than the previously mentioned modulations. However, since QAM presents a variant envelope, it will be necessary to consider linear power amplifiers which are less efficient than the classical class-C power amplifiers used with the current FM-type modulations.

In order to achieve both spectrum and power efficiency, several techniques of linearizing power amplifiers are proposed here. These techniques can be categorized into four types: feed-forward (1), feedback (2), predistortion (3), and LINC transmitter (4). Among them, one of the most promising is the LINC — an acronym for *LI*near amplification using *N*on-linear *C*omponents. The LINC transmitter, like the feed-forward type, does not use a feedback loop. It differs from that implementation since it also does not use signal error to compensate for the distortion introduced by the power amplifier. The basic principle of the LINC transmitter is to represent any arbitrary bandpass signal, which may have both amplitude and phase variations, by means of two signals which are of constant amplitude and only have phase variations. These two angle modulated signals can be amplified separately using efficient high power non-linear devices. Finally, the amplified signals are passively combined to produce an amplitude modulated signal.

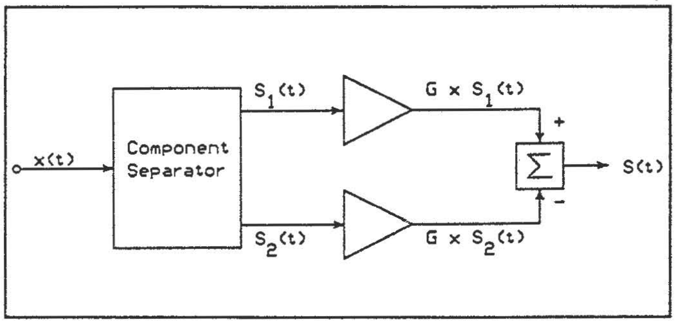

Figure 1. The LINC transmitter.

Although the LINC principle was developed during the 1930's to improve the efficiency and linearity of the AM-broadcast transmitter, it has not been widely used due to the complicated structure needed for generating the two angle modulated signals when analog circuits are used. However, the advent of Digital Signal Processing (DSP) devices permits complicated operations in a simple manner. The technique of using DSP to generate two angle modulated signals is called the component separator.

Even though some work has been published in this respect (5), little effort has been made toward characterizing the impact of these circuits' misfunction or imbalance on system performance. This article considers errors due to both the digital signal processing and the RF processing. A classical two tone test is considered to characterize the system performances.

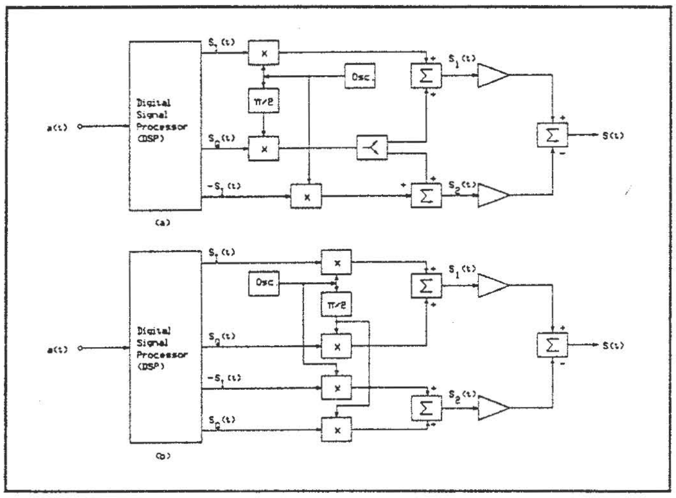

Figure 2. The LINC transmitter with DSP devices.

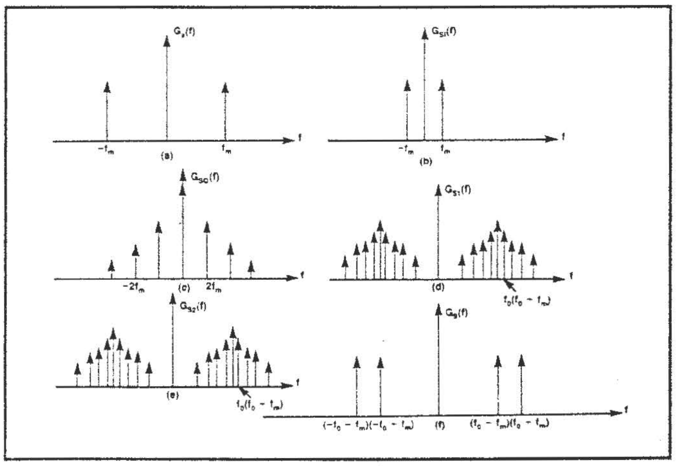

Figure 3. Power spectrum plots of (a) the input signal a(t); (b) $S_I(t)$;(c) $S_Q(t)$; (d)$S_1(t)$; (e) $S_2(t)$; (f) S(t).

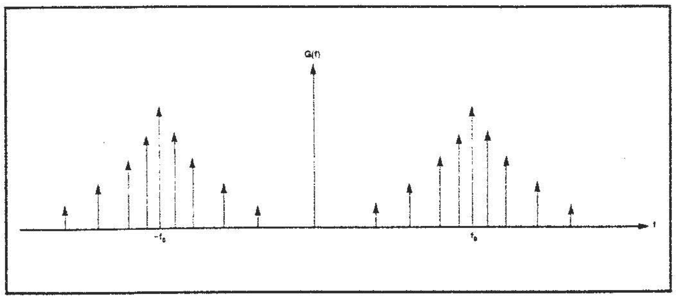

Figure 4. Power spectrum of the interfering signal.

System Description

Figure 1 shows the schematic drawing of LINC system, where:

$S(t) = G[a(t)\cos(\omega_0 t + \phi)]$

$S_1(t) = V/2\ [\sin(\omega_0 t + \phi + \psi(t)]$

$S_2(t) = V/2\ [\sin(\omega_0 t + \phi - \psi(t)]$

with $\psi(t) = \sin^{-1}(a(t)/V)$, and $\max(a(t)) \leqslant V$. Obviously, the component separator is a non-linear device because it uses two phase modulations with inverse sine characteristics. In the past, several structures were proposed to obtain such characteristics (6,7). These are all very complicated implementations of the LINC structure and have not been widely used.

However, considering that the signals $S_1(t)$ and $S_2(t)$ can be expressed as:

$$S_1(t) = S_I(t)\cos(\omega_0 t + \phi) + S_Q(t)\sin(\omega_0 t + \phi)$$
$$S_2(t) = -S_I(t)\cos(\omega_0 t + \phi) + S_Q(t)\sin(\omega_0 t + \phi)$$

with:

$$S_I(t) = a(t)/2$$
$$S_Q(t) = \frac{1}{2}\sqrt{V^2 - a(t)^2}$$

$S_1(t)$ and $S_2(t)$ can be generated using the circuit shown in Figure 2(a). In this scheme, the only non-linear operation is done when $S_Q(t)$ is generated. As this depends on a(t), a baseband signal, the non-linear operation root-square could be easily obtained using a DSP device. Then, only a DSP device, three balanced modulators, two combiners and a power splitter need to be used for generating $S_1(t)$ and $S_2(t)$. An equivalent scheme that uses four balanced modulators and two combiners is shown in Figure 2(b). In order to understand the behavior of the LINC transmitter (Figure 3), the power spectrum of all the signals present in the different points of the scheme is shown in Figure 2(b).

System Impairments

In a practical LINC amplifier there are two main mechanisms that degrade the overall performance:

- Errors due to the digital signal processing that produce imperfect generation of the constant amplitude phase modulated signal component, $S_1(t)$ and $S_2(t)$.
- Errors due to the RF processing that cause amplitude and delay imbalances between the parallel signal paths.

The errors introduced in the amplitude and/or phase of the $S_1(t)$ and $S_2(t)$ could be caused by the quantification error, due to the limited length of the words used in the DSP and the imperfect implementation of the square law characteristic. Nowadays, the available DSP chips use 16 bits/word or even longer ones (8) that make quantification errors negligible. Moreover, the use of appropriate algorithms allows square operation with a high degree of precision. So, in general, the effects of this kind of impairment are negligible.

With respect to the errors due to the RF processing, the following subjects should be considered: imbalance in gain between the two paths; imbalance in delay between the two paths; and imbalance in the nonlinear characteristics of power amplifiers.

Imbalance in Gain Between the Two Paths: In this case,

$$S(t) = GS_1(t) - (G + \Delta G)S_2(t)$$
$$= Ga(t)\cos(\omega_0 t + \phi) - \Delta G\,\frac{V}{2}\sin[\omega_0 t + \phi - \psi(t)]$$

A residual level of one of the phase modulated signals appears in the system output. In order to analyze the impact of this undesired signal on the linearity of the system, consider the classical two tone test. In this case, considering that $\phi = 0$ (with no loss in generality) the useful output signal is:

$$S_U(t) = G[A\cos(\omega_1 t) + A\cos(\omega_2 t)] = G[a(t)\cos(\omega_0 t)]$$

with $\omega_0 = (\omega_1 + \omega_2)/2$ and,

$$a(t) = 2A\cos(\omega_m t) = 2A\cos[(\omega_1 - \omega_2)t/2]$$

The interfering signal is:

$$S_I(t) = \Delta G\,\frac{V}{2}\sin[\omega_0 t - \psi(t)]$$
$$= -\Delta G A\cos(\omega_m t)\cos(\omega_0 t) + \frac{\Delta G}{2}\sqrt{V^2 - [2A\cos(\omega_m t)]^2}\sin(\omega_0 t)$$

In the above expression, the first part adds directly to the useful signal without producing distortion, but the second part generates a set of interfering spectral lines inside and outside the desired bandwidth.

The most important of the interfering spectral lines, as shown in Figure 4, is placed at ω_0 whose level is given by:

$$IL = \frac{2}{\Pi}\frac{\Delta G}{2} V\int_0^{\pi/2}\sqrt{1 - A_x^2\cos^2(\vartheta)}\,d\vartheta = \frac{2}{\Pi}\frac{\Delta G}{2} VE(A_x,\pi/2)$$

with $A_x = 2A/V \leqslant -1$, and $\vartheta = \omega_0 t$, and E(x,y) is the elliptic integral of the second kind (9).

Considering that the level of the useful spectral line is $A(G + \Delta G/2) = AG$, the undesired response rejection at the output is defined as:

$$U_R(\text{dB}) = -20\log\left[\frac{\Delta G}{G}\right] - 20\log\left(\frac{\frac{2}{\Pi}E(A_x,\pi/2)}{A_x}\right)$$

Figure 5 shows the evolution of the output rejection against the $\Delta G/G$ value with parameter A_x. It is important to emphasize that the undesired response rejection at the output depends not only on the $\Delta G/G$ value but also on the relative level of the input modulating signal 2A/V. The smaller the relative level of the input level (A_x), the more important the degradation of the system performance.

For convenience insure that the input signal is as close as possible to the maximum permissible input level.

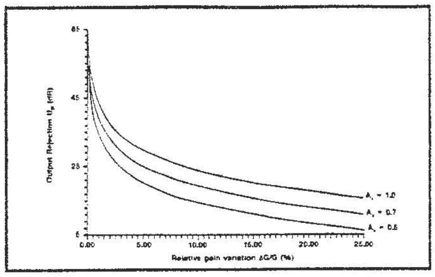

Figure 5. Output rejection versus relative gain variation.

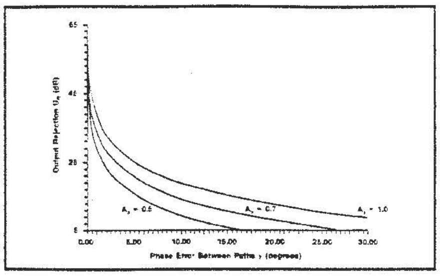

Figure 6. Output rejection versus phase imbalance.

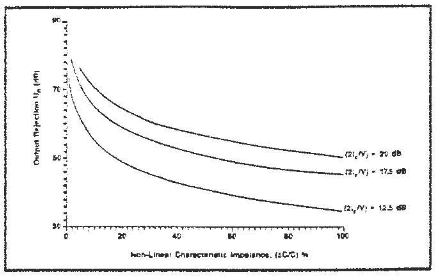

Figure 7. Output rejection versus relative imbalance.

Note that a small gain difference between both paths produces a significant degradation. For example, if the gain imbalance between paths is about 10 percent and the relative input signal is equal to one, the output rejection decreases to 19 dB, but when the relative input signal is only 0.5 then the output rejection is as low as 9 dB. Even though, the gain imbalance produces important degradations on system performances, in practice, this effect can be easily overcome by an appropriate adjustment of the amplifier gains.

Imbalance in the Delay Between Two Paths: If the two path signals have two different delays in the output of the combiner, the signals don't combine in phase, and it results in a high degree of distortion. Therefore,

$$S(t) = GS_1(t) - GS_2(t - \tau)$$

where τ can be expressed as γ/ω_0. After some algebraic operations, the above expression can be related as:

$$S(t) = Ga(t)\cos(\omega_0 t + \phi)\cos(\gamma) + G(1 - \cos(\gamma))S_2(t) + G\sin(\gamma)S_3(t)$$

with,

$$S_3(t) = (V/2)\cos[\omega_0 t + \phi - \psi(t)] = S_I(t)\sin(\omega_0 t + \phi) + S_Q(t)\cos(\omega_0 t + \phi)$$

Comparing the expressions for $S_2(t)$ and $S_3(t)$, it is obvious that both are phase modulated signals whose modulus power spectrums are equal. So, with $\gamma << 1$.

$$S(t) \cong Ga(t)\cos(\omega t + \phi) - G\gamma S_3(t)$$

As before, assuming that $a(t) = 2A\cos(\omega_m t)$, and considering that $S_3(t)$ is a phase modulated signal, the most relevant of all the interfering spectral lines is placed at ω_0. Its level is given by:

$$IL = \frac{G\gamma V}{\Pi} E(A_x, \pi/2)$$

with $A_x = 2A/V$. The undesired response rejection at the output can now be expressed as:

$$U_R(dB) = -20\log(\gamma) - 20\log\left[\frac{\frac{2}{\Pi}E(A_x, \pi/2)}{A_x}\right]$$

Figure 6 shows the evolution of the undesired response rejection at the output against the phase imbalance γ. Observe that the small delay imbalance between both paths causes a high degree of non-linear distortion. For example, a phase error between both paths as low as 2.5 degrees diminishes the output rejection to only 31 dB. For the most favorable value of A_x, A_x is equal to 1. In fact, this is the most important cause of the signal distortion of the LINC transmitter. As a result, a recent modification in the LINC structure which permits the control of phase balance in both paths, has been proposed in reference 10.

Imbalance in the Non-Linear Characteristic of Both Power Amplifiers: The non-linear behavior of the power amplifiers can be characterized by means of the following expression:

$$v_0(t) = gv_i(t) + bv_i^2(t) + cv_i^3(t)$$

where g is the gain for low signal level, b and c denote the second and third order coefficients of the non-linear characteristic respectively and, in this case, the input signal, $v_i(t)$, is $S_1(t)$ or $S_2(t)$. Then, defining $v_{01}(t)$ (and $v_{02}(t)$) as the signal at the output of the power amplifier,

$$v_{01}(t) = g\left[1 + \frac{3}{4}\frac{c_1}{g}\left(\frac{V}{2}\right)^2\right]S_1(t) + 2f_0 \text{ terms} + 3f_0 \text{ terms}$$

$$v_{02}(t) = g\left[1 + \frac{3}{4}\frac{c_2}{g}\left(\frac{V}{2}\right)^2\right]S_2(t) + 2f_0 \text{ terms} + 3f_0 \text{ terms}$$

In the above expressions c_1 (or c_2) denotes the third order characteristic of the first (and second) non-linear power amplifier. The gain for low signal level g has been assumed equal in both amplifiers. In the presence of a phase modulated input signal, the effect of the non-linear characteristic of the amplifier is to compress the power gain. Obviously, this means that c_1 (c_2) is a negative constant. The terms centered in $2f_0$ and $3f_0$ will be eliminated by the filters and matching circuits placed at the output of the power amplifier.

From $v_{01}(t)$ and $v_{02}(t)$, the output signal S(t) can be written as:

$$S(t) = v_{01}(t) - v_{02}(t) = g\left[1 + \frac{3}{4}\frac{c_1}{g}\left(\frac{V}{2}\right)^2\right]a(t)\cos(\omega_0 t + \phi) + \frac{3}{4}\left(\frac{V}{2}\right)^3 \Delta c\,\sin[\omega_0 t + \phi - \psi(t)]$$

where $\Delta c = c_1 - c_2$. The second term shows the effect of the imbalance between the non-linear power amplifiers.

In order to analyze the impact of this interfering signal on the linearity of the system, consider a classical two tones test. In this case, as shown before, the interfering signal generates a set of undesired spectral lines, the most important of which is placed about ω_0. This line has a level given by:

$$IL = \frac{2}{\Pi}\left[\frac{3}{4}\left(\frac{V}{2}\right)^3 \Delta c\right] E(A_x, \pi/2)$$

Next, considering that the level of the useful spectral line is:

$$Ag\left[1 + \frac{3\ c_1}{4\ g}\left(\frac{V}{2}\right)^2\right] \cong Ag$$

and after some algebraic effort, the output rejection to the undesired response can be defined as:

$$U_R\,(dB) = -20 \log\left[\frac{\Delta c}{c}\right] - 20 \log\left(\frac{\frac{2}{\Pi} E(A_x, \pi/2)}{A_x}\right) - 40 \log\left[\frac{V}{2I_p}\right]$$

Figure 7 shows a plot of the undesired response rejection at the output against the relative imbalance between both non-linear characteristics ($\Delta c/c$). The ratio between the third order intercept point (I_p) and the level of the phase modulated signal (V/2) is a constant. In all the cases considered, A_x is equal to one. For high values of the I_p/V ratio, the system is very insensitive to the relative variations of the non-linear characteristic, $\Delta c/c$. The more non-linear the power amplifiers are, the more sensitive the LINC transmitter is to the imbalance. However, as in Figure 7, for relative errors with $\Delta c/c$ as high as 50 percent, the output rejection only decreases to 57 dB for $2\,I_p/V$ = 20 dB and to 52 dB for $2\,I_p/V$ = 17.5 dB.

The effect of both digital and RF signal processing in a LINC transmitter was analyzed. The errors due to the digital signal processor have a negligible effect on the system performances. However, the errors due to the RF processing have more effects and need to be considered. This article distinguished between three path imbalances. The imbalance in the gain of both the power amplifiers produces a significant reduction in the linearity of the LINC transmitter.

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