SSPA Linearization

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ABSTRACT

Many new communications services require highly linear high power amplifiers (HPAs). This paper is an extension of an earlier paper, which discussed the merits of linearization and its application to Traveling Wave Tube Amplifiers (TWTAs). In this paper the emphasis is on solid state power amplifiers, SSPAs. Various methods of linearization are compared and found to offer significant benefit when used with both bipolar and Si/GaAs FET power amplifiers. Problems unique to SSPAs are highlighted. Predistortion is shown to be the preferred linearization approach for many applications.

INTRODUCTION

The growth of the cellular telephone service and other forms of personal communications have created a demand for highly linear amplifiers. In the case of cellular telephony, it is often more convenient and cost effective to transmit several carriers through a common amplifier rather than to use multiple amplifiers and a lossy multiplexer, as is illustrated in Figure 1. To avoid unacceptably high intermodulation distortion (IMD), the common amplifier must be highly linear.

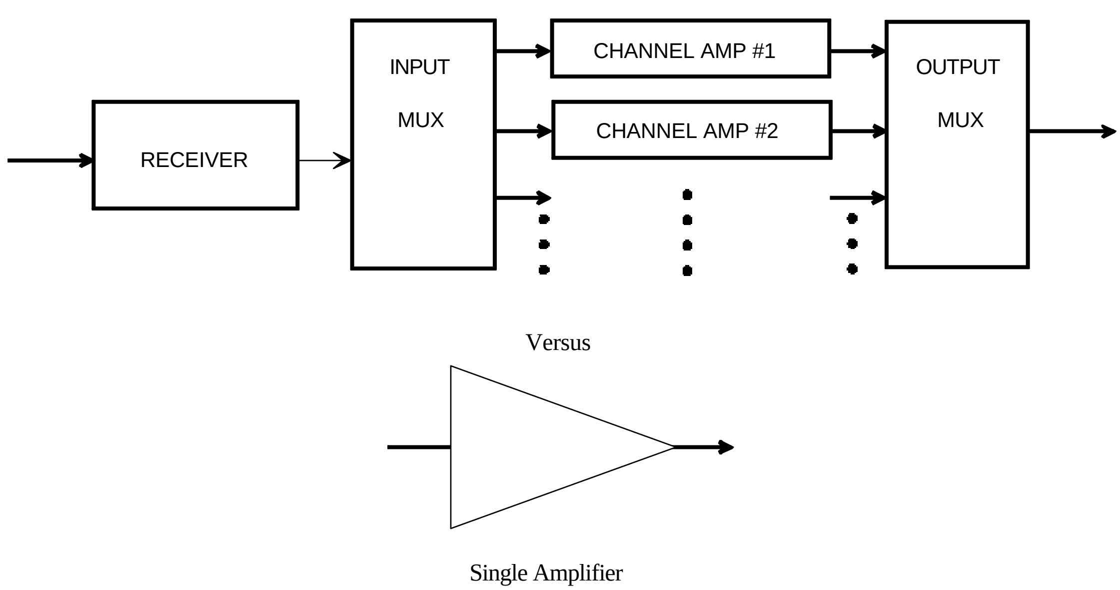

Figure 1. In cellular telephony transmitting several carriers through a common amplifier is more cost effective than using multiple amplifiers.

For the transmission of a single carrier, IMD is usually not a limitation. However with digitally modulated signals, spectral regrowth (SR) can be a serious problem, and manifests itself in a form equivalent to IMD. SR is not unique to digital signals, but an aspect of angle modulation (FM and PM). Angle modulated signals have a theoretically infinite bandwidth; for example, the spectrum of a sinusoidal modulated FM signal,

$$Ac\cos(\omega ct + M\cos[\omega mt]) = Ac\sum_{n=-\infty}^{n=\infty} J_n(M)\cos([\omega c+n\omega m]t) \quad (1)$$

contains an infinite number of sidebands. In practice the bandwidth is limited to a finite frequency band, beyond which sideband amplitude drops off rapidly. Analogue-FM has an approximate bandwidth given by Carson's rule.[2]

$$BW = 2\,(\Delta f + fm) \quad (2)$$

where Δf is the peak deviation and fm is the modulation frequency. The effective bandwidth of angle modulated digital signals can be much greater than predicted by (2), due to the high frequency components of the modulating waveform. To reduce their bandwidth to a more acceptable value, digital waveforms are normally low-pass filtered before modulation. Because of the mechanics of most digital modulators, (which are not true angle modulators), the amplitude of the carrier is also modulated by this process. In addition any "band limiting" (filtering) of an angle modulated signal will introduce amplitude modulation. It is primarily this incidental amplitude modulation which causes the SR when a digital signal is passed through a non-linear amplifier. The distortion of the induced amplitude waveform produces IMD products, which increase the signal's spectrum.

In practical amplifiers there is usually a substantial change in signal phase angle with power level.

$$\theta = f(Pin) \quad (3)$$

This change in phase with amplitude converts the variations in signal level to angle modulation sidebands. These new sidebands further broaden the signal bandwidth. Amplitude and phase induced spectral products add vectorally and are classified in general as IMD.

SR is a major concern in personal communications since transmission often occurs on a channel adjacent to one in which reception of a much weaker distant signal may be taking place. To ensure freedom from interference, transmitter IMD products must be below the carrier (C/I ratio) by anywhere from 35 to greater than 65 dB, depending on the application.[3]

The summation of the IMD terms in the adjacent channel is referred to as the adjacent channel power (ACP).

$$ACP = \Sigma\, IMDs \Big|_{\text{in the adjacent channel}} \quad (4)$$

These levels of linearity are considerably higher than had been required of communications amplifiers in the past, except for some very special applications.

SATURATED POWER

All amplifiers have some maximum output power capacity, referred to as *saturated power* or simply *saturation* (SAT) - see Figure 2. Many SSPAs are sensitive to overdrive and can be easily damaged by operation at or beyond saturation. In addition, SSPAs tend to approach saturation exponentially. These factors make engineers reluctant to measure and use saturated power as a reference for comparison of SSPA performance. They prefer to use the power at which an amplifier's gain compresses by 1 dB as the reference (REF) for amplifier comparison.

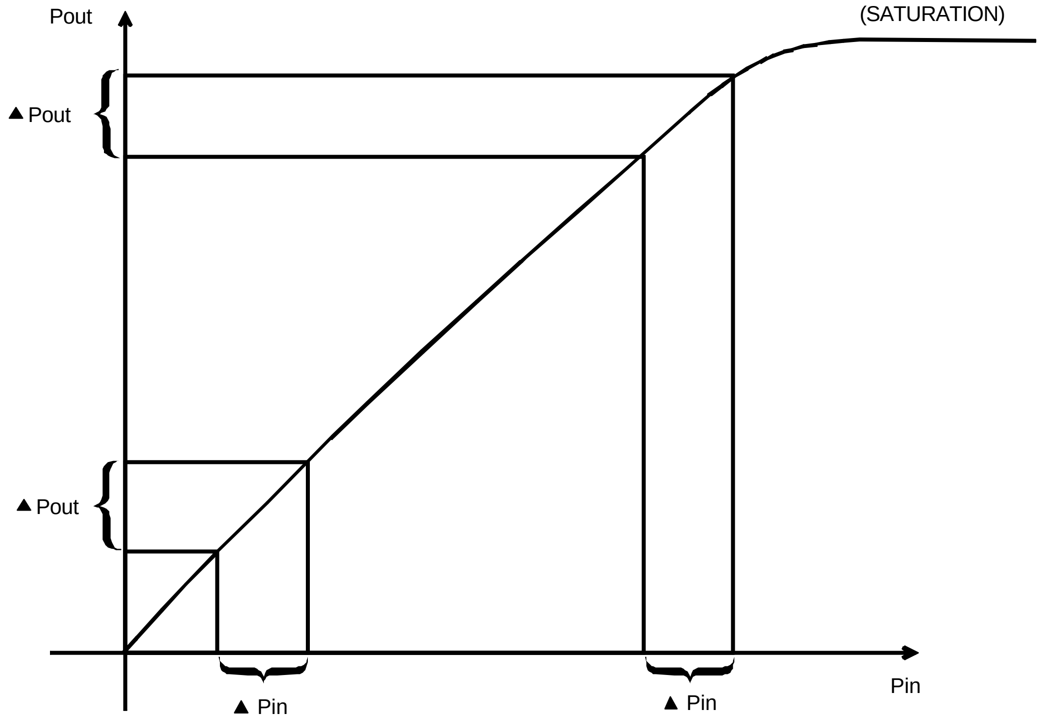

Figure 2. Saturated output power is approached exponentially by a typical Class A SSPA.

$$\text{REF} = 1\ \text{dB CP} = \text{SAT} - \text{D} \qquad (5)$$

For SSPAs with reasonable linearity, the difference (D) in output level between SAT and the 1 dB compression point (CP) is about 1 dB. Unfortunately D varies from amplifier to amplifier. Amplifiers with high linearity will have a smaller difference ($D < .25$ dB), while amplifiers with poor linearity can have a difference of several dB ($D > 2$ dB). For this reason, in this paper, relative amplifier performance will be referenced to (single carrier) saturated power. Output power backoff (OPBO) will be relative to an amplifier's single carrier saturated power. (For most SSPAs, SAT can be safely determined using a network analyzer in a rapid power sweep mode. For amplifiers that are especially thermal sensitive, pulsed power sweep techniques may be used.)[4] **When comparing the data presented here with that of SSPAs based on a 1 dB CP reference, an appropriate correction factor should be assumed.**

Generally an SSPA's greatest efficiency will occur at or near saturation. Similarly the closer to saturation a *linear* amplifier, (class A and to a large extent class AB), is driven, the greater the amount of distortion it produces. This means that typically a (class A) SSPA will have to be backed-off about 5.5 dB (for a C/I = 35 dB), and by more than 15 dB (for a C/I = 65 dB) to satisfy cellular/PC adjacent channel IMD requirements. These are huge reductions in usable output power, and apply for both FET and bipolar devices.

LINEARIZATION

Linearization is a systematic procedure for reducing an amplifier's distortion. There are many different ways of linearizing an amplifier. Usually extra components are added to the design of a conventional amplifier. Often these extra components can be configured into a subassembly or *box* that is referred to as a linearizer. Linearization allows an amplifier to produce more output power and operate at a higher level of efficiency for a given level of distortion. Feed-forward, Feedback and Predistortion are some common forms of linearization.

Feed-Forward Linearization

For SSPAs, Feed-forward (FF) techniques have received much attention in recent years. A block diagram of a basic FF system is shown in Figure 3. This system consists of two loops. The first loop subtracts samples of the input signal (Sin) from the output signal (Sout1) to produce a sample of the main amplifier's distortion. Sout1 consists of the amplified input signal plus any distortion introduced by the amplifier (IMD).

$$\text{A0 SSin} \angle \Phi 0 = -\text{SSout1}$$

or

$$\text{A0 K0 Sin} \angle \Phi 0 = \text{G K1 Sin} \angle (\Phi\text{amp} + 180°) \quad (7)$$

then Sin is canceled and the output of loop 1 is K1 IMD. A0 and Φ0 are respectively the attenuation and phase shift introduced in loop 1 for adjustment of the carrier cancellation.

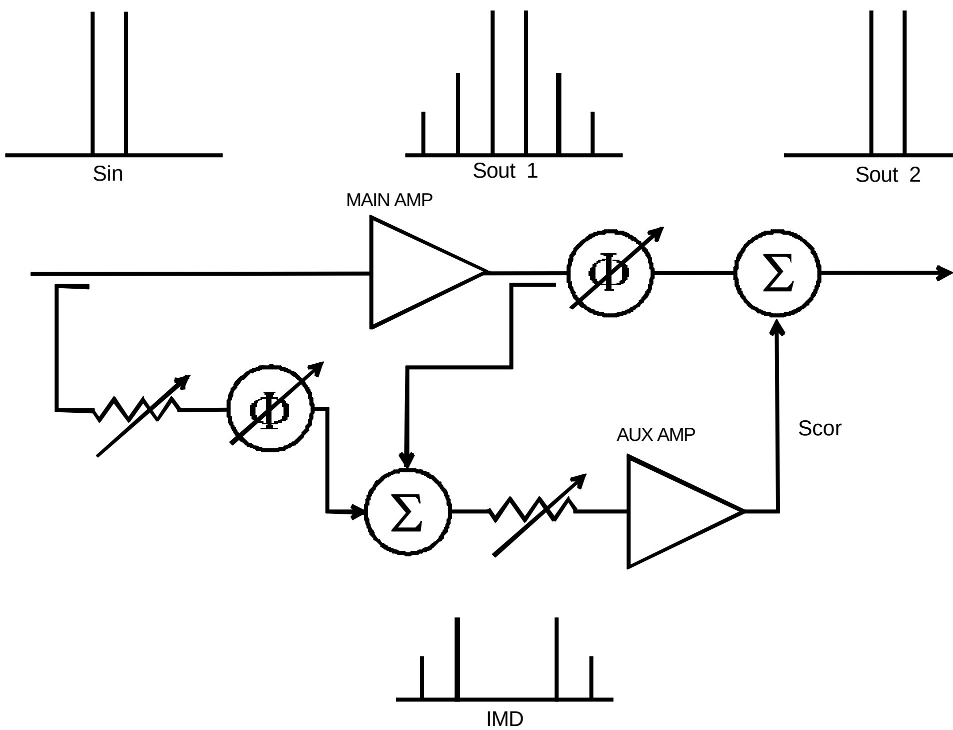

Figure 3. Feed-forward linearization employs two loops for the cancellation of IMD.

$$\text{Sout1} = \text{G Sin} \angle \Phi\text{amp} + \text{IMD} \quad (6)$$

G is the gain and ∠Φamp is the phase shift introduced by the main amplifier. The samples of Sin (SSin) and Sout1 (SSout1) are respectively

$$\text{SSin} = \text{K0 Sin and SSout1} = \text{K1 Sout1}$$

K0 and K1 are the coupling coefficients of the directional couplers used to sample Sin and Sout1 respectively. If SSin is attenuated and delayed in phase such that

The second loop subtracts the amplified sampled distortion of loop 1 from a delayed Sout1 to produce *ideally* a distortion free output signal (Sout2). The loop 1 output signal is amplified by an auxiliary (aux) amplifier of gain GA and phase shift Φaux to provide a correction signal (Scor) of sufficient level to cancel the distortion introduced by the main amplifier. Scor is combined with the main amplifier signal at a final directional coupler of coefficient K2.

If

$$Scor = A1\ GA\ K1\ K2\ IMD \angle(\Phi aux + \Phi 1) = IMD \angle(\Phi m + 180°) \quad (8)$$

then the SSPA output will be distortion free. A1 and Φ1 are respectively the attenuation and phase shift introduced in loop 2 for adjustment of the distortion cancellation. Φm is a delay added after the main amplifier to equalize the delay introduced by the aux amplifier.

$$Sout2 = Sout1 \angle \Phi m + Scor \quad (9)$$

From this discussion it may appear that undistorted output can be obtained from a FF amplifier right up to SAT. Saturated output power can never be obtained from a FF amplifier because of the losses in the phase shifter and couplers, which must be located after the main amplifier.

The main signal, Sout1, is reduced in amplitude by a factor (R1) due to passing through the K1 coupler.

In dB

$$R1 = 10\ LOG(1 - 10^{-(k1/10)}) + L1 \quad (10)$$

where L1 is the dissipation loss of the coupler in dB. K1 can be made very small, provided the main amplifier has sufficient gain. (A K1 of -30 dB is not unusual). The K2 of the final directional coupler must also be relatively small to minimize the loss of output power (R2). Since the two signals, (carriers and distortion), being combined are not at the same frequency, power will be split between the load and the coupler's *dump* port. The R2 power loss in dB as function of K2 is described by equation (10) with 2 substituted for 1 in the variable names. The overall loss in saturated power (ΔSAT) is

$$\Delta SAT = 10\ LOG(1 - 10^{-(k1/10)}) + 10\ LOG(1 - 10^{-(k2/10)}) + L1 + L2 + Lm \quad (11)$$

where Lm is the loss of the delay line (Φm). In practice it is very difficult to achieve a ΔSAT of less than 1 dB.

ΔSAT can be considered the minimum OPBO of a FF amplifier. In actuality, ΔSAT must be added to the difference between the saturated power of an amplifier with single and multi-carrier signals.[1] This factor can vary from about .5 to >1 dB for SSPAs. Further more, the amplifier's *true* SAT power should not be considered only that of the main amplifier. A FF amplifier combines both the power of the main and the aux amplifier. The sum of the saturated power of both these amplifiers should be considered when comparing the relative OPBO performance of different methods of linearization.

Practical considerations limit the size of the aux amplifier. This limits Scor and in turn the undistorted FF output level. The smaller K2 is set, the larger in power the aux amplifier must be sized. The aux amplifier must also be operated relatively linear so as not to *distort* the distortion signal, and thus introduce distortion of its own. Figure 4 shows the relationship between minimum OPBO (from single carrier SAT of the main amplifier) and size of the aux amplifier (relative to the main amplifier), for cancellation of IMD. Minimum OPBO is given for different values of output coupler coefficient K2. These results depend on the linearity of the main and aux amplifiers and on the resistive loss of the couplers and the delay line. Linear characteristics typical of a class A GaAs FET SSPA were assumed for both amplifiers, and resistive losses of 1 dB were assumed for the passive output components. Figure 4 shows that with an aux amplifier of size equal to the main amplifier (0 dB), cancellation of IMD can be achieved up to only -4.2 dB from single carrier SAT.

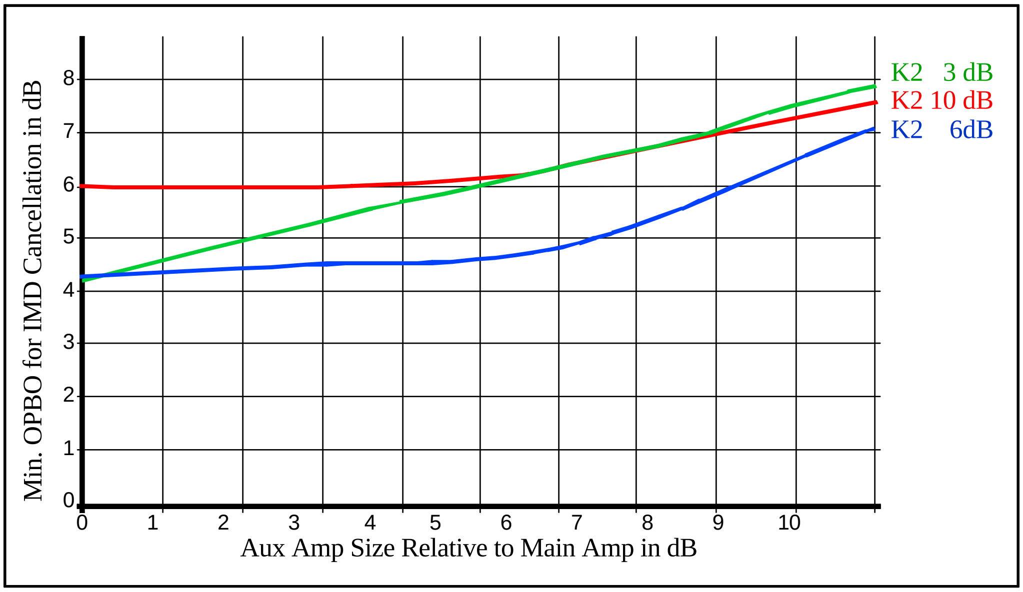

Figure 4. The mnimum OPBO for cancellation of IMD by a feed-forward amplifier depends on auxiliary amplifier size and output coupler coefficient.

If the combined saturated output power levels of the main and aux amplifiers are used as the reference for OPBO, then IMD cancellation is limited to OPBOs of less than -6.3 dB. The relationship between minimum OPBO (from single carrier SAT of the main plus aux amplifiers) and size of the aux amplifier is shown in Figure 5. When overall amplifier power capacity is considered, the minimum *corrected* OPBO occurs for an aux amplifier sized approximately 3 dB smaller than the main amplifier, and a K2 of about 6 dB. (A minimum IMD cancellation of 20 dB was assumed. If only 10 dB of cancellation is acceptable, an additional 1 to 2 dB increase in output level can be achieved.) In practice other factors limit IMD reduction and perfect cancellation can never be achieved.[5,6,7]

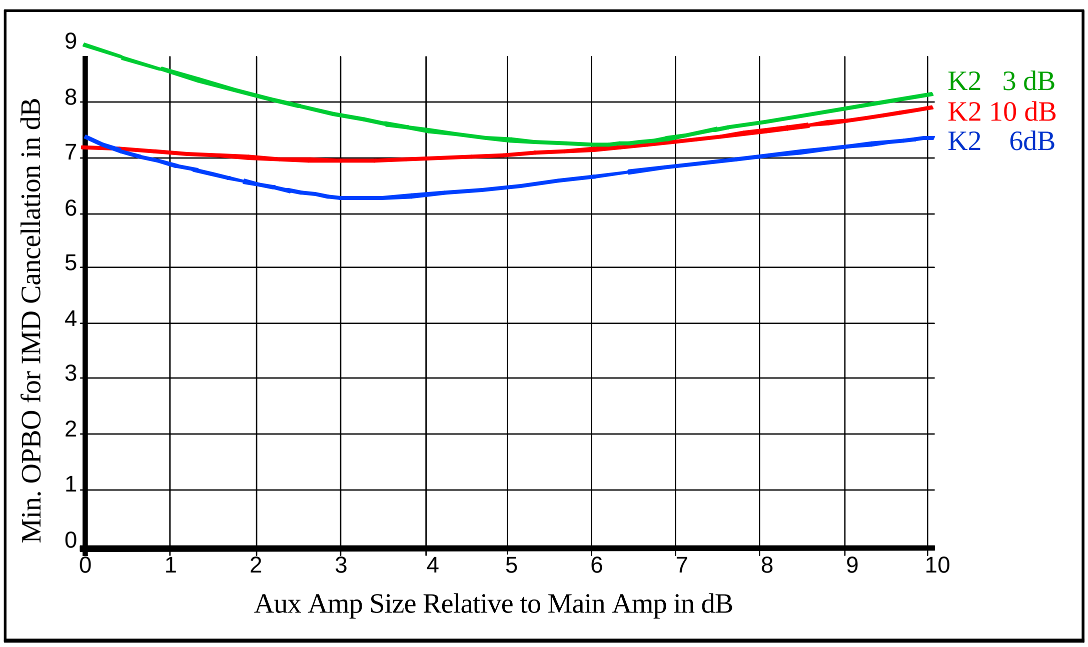

Figure 5. Additional OPBO is required for cancellation of IMD when the output power of the auxiliary amplifier and the main amplifier are used as the reference.

Figures 4 and 5 reveal why FF is not a good choice for linearization of amplifiers near SAT. Other linearization methods can provide superior IMD cancellation with considerably less complexity. However FF becomes competitive for OPBOs greater than 6 ~ 7 dB, and for high linearity requirements may be the system of choice.

Feedback Linearization

There has been considerable work on feedback for the linearization of RF and microwave amplifiers. Feedback techniques can be divided into several distinct branches. The use of linear networks for feedback is well documented, but has seen little application at microwave frequencies.[8] The reason for this reluctance is probably concern with amplifier stability and the difficulty in making networks with non-ideal components function over wide frequency bands.

$$Ve = |DSout - DSin| \tag{12}$$

where DSout and DSin are respectively the detected output and input signals. Ve can be used to control the gain of the amplifier by means of a voltage variable attenuator. Superior linearity is obtained by correcting both gain and phase. The magnitude and phase of Ve can be determined as illustrated in Figure 6. The resulting error signals are used to respectively control an attenuator and a phase shifter so as to minimize the error.

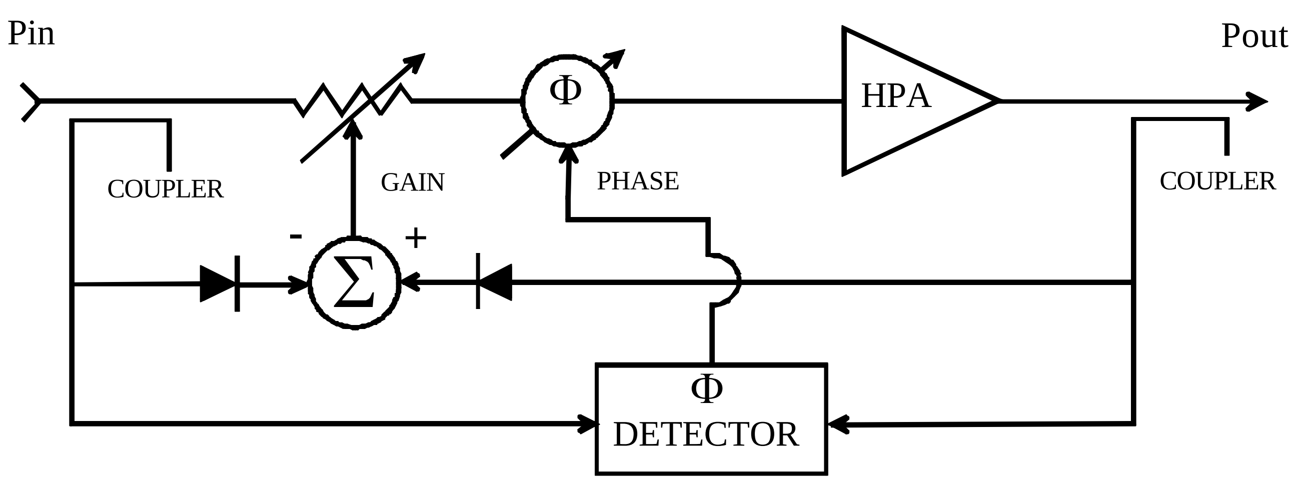

Figure 6. The difference in magnitude and phase between input and output signals is detected and the error signals used to control amplifier gain and phase.

Indirect feedback (IFB) techniques have been more widely applied.[9,10] In these approaches an amplifier's input and output signals are detected, and the resulting *baseband* signals compared. The error signal (Ve) is used to modify the amplifier's characteristics so as to minimize distortion.

A variation of IFB, know as Cartesian feedback, separates the signal into in-phase and quadrature components. This eliminates the need for the phase detectors and shifters, and still allows the correction of gain and phase by adjusting the amplitudes of two orthogonal components. Figure 7 shows an example of a Cartesian feedback system in which the input and output signals are separated into orthogonal components and detected. The resulting baseband in-phase and quadrature components are compared and used to control the attenuators in a vector modulator. Cartesian feedback is most often used with QPSK modulation. In this case the detected components are subtracted from the in-phase and quadrature modulation signals as shown in Figure 8.

Today correction is done at baseband using digital signal processing techniques. Very high linearity can be achieved with IFB. The resulting system is self-correcting for changes due to environmental and aging effects. IFB's principal limitation is an inability to handle wideband signals. It is "practically" very difficult to make a feedback system responds to signal envelope changes much greater than a few MHz due to the inherent time delay built into the system.[10]

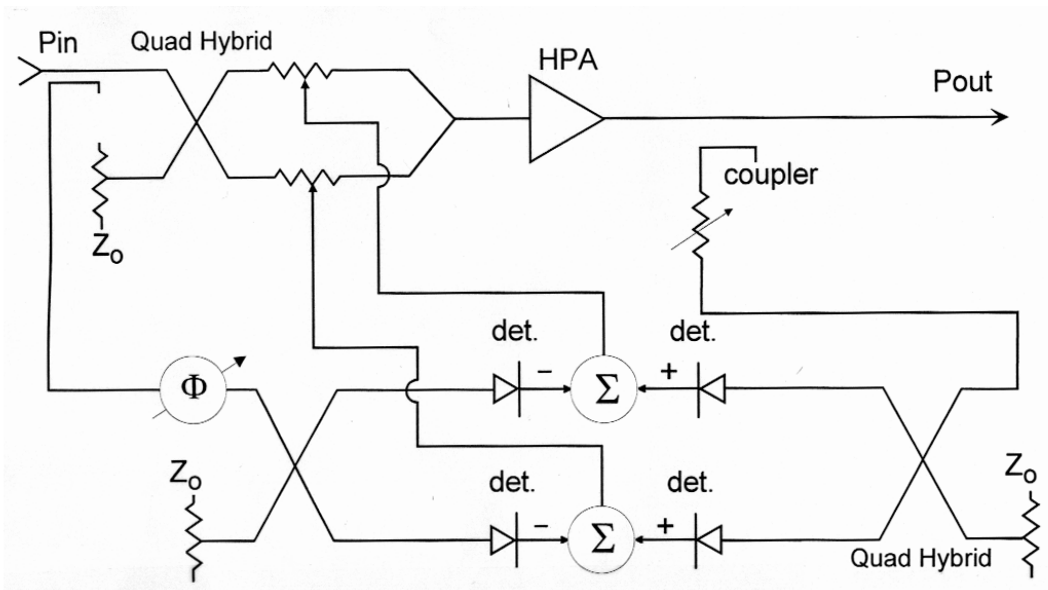

Figure 7. Cartesian feedback eliminated the need for phase correction by using the difference between in-phase and quadrature signals to control attenuators in a vector modulator.

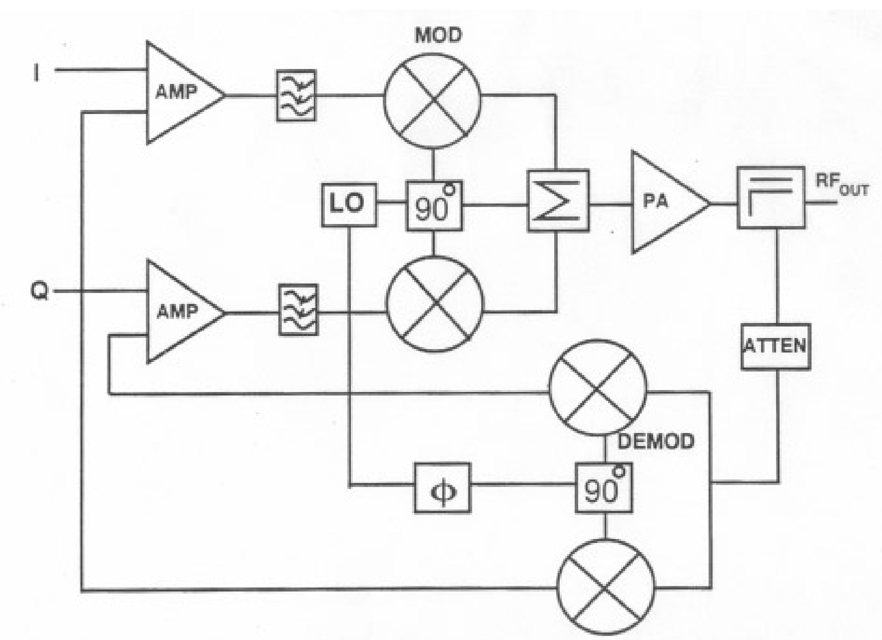

Figure 8.
Cartesian feedback is used with QPSK modulation to directly correct the in-phase and quadrature modulation signals.

Schemes in which amplifier bias is changed in response to Ve have been referred to as Dynamic Electronic Bias Systems (DEBS). However, most DEBS use the input signal as the reference without comparison to produce an indirect form of FF linearization.[11]

Predistortion Linearization

Predistortion (PD) linearizers have been used extensively in microwave and satellite applications because of their relative simplicity, and their ability to be added to existing amplifiers as separate stand alone units. PD linearizers, unlike FF, can provide a viable improvement in linearity near saturation, but can be difficult to apply in applications requiring very high linearity (C/I > 50 dB). PD linearizers generate a non-linear transfer characteristic, which is the reverse of the amplifier's transfer characteristics in both magnitude and phase – see Figure 9. An alternate way of thinking of a PD linearizer is to view the linearizer as a generator of IMD products. If the IMDs produced by the linearizer are made equal in amplitude and 180 degrees out phase with the IMDs generated by the amplifier, the IMDs will cancel. This condition occurs when the gain and phase of the linearized amplifier remain constant with change in power level.

In dB, the gain of the linearizer (GL) must increase by the same amount the amplifier's gain (GA) decreases.

$$GL(P_{outL}) - GL_{ss} = -\left[GA(P_{inA}) - GA_{ss}\right] \Big|_{P_{outL} = P_{inA}} \quad (13)$$

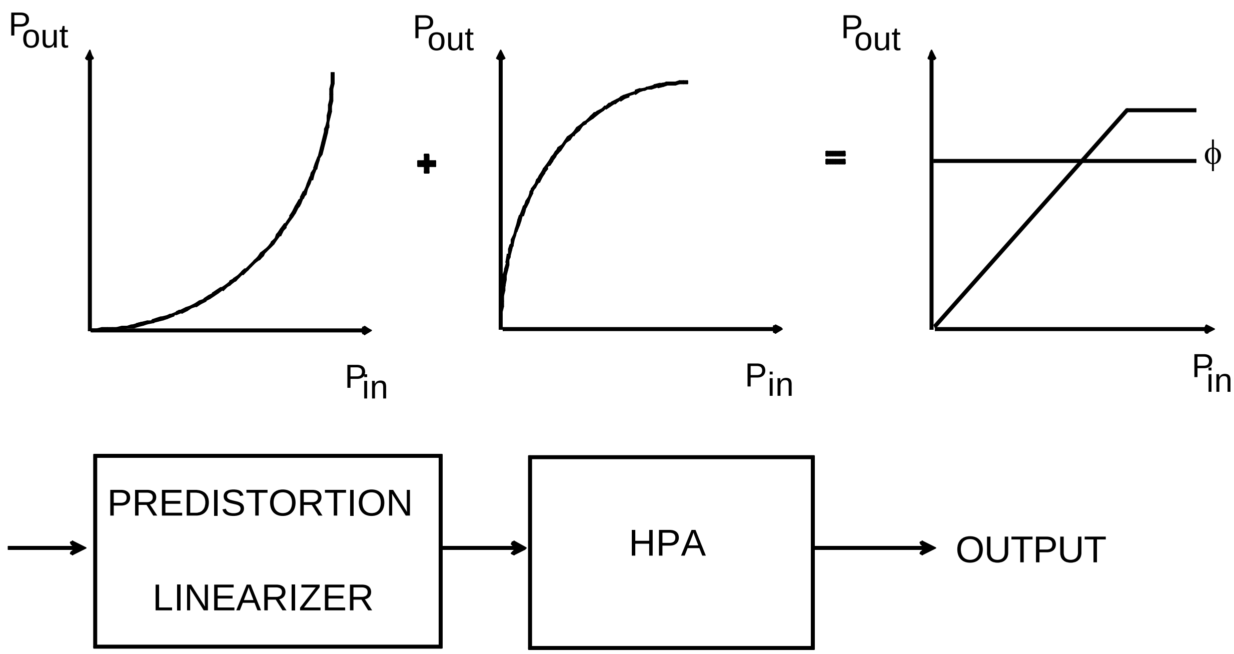

Figure 9. Ideally the gain and phase of a linearized amplifier should remain constant up to saturation.

where GL_{ss} and GA_{ss} are respectively the small signal gains of the linearizer and the amplifier, and $GL(P_{outL})$ and $GA(P_{inA})$ are respectively these gains as a function of linearizer output and amplifier input levels. Likewise the phase shift introduced by the linearizer must increase by the same amount the amplifier's phase decreases, (or vice-verse depending on the direction of phase change by the amplifier).

$$\Phi L(P_{outL}) - \Phi L_{ss} = -[\Phi A(P_{inA}) - \Phi A_{ss}] \Big|_{P_{outL} = P_{inA}} \quad (14)$$

When these conditions are met, the result is the composite linear transfer characteristic shown in Figure 9. This is the response of a so-called *ideal limiter*. Once an amplifier has saturated, it is impossible to obtain more output power by driving the amplifier harder. Thus the best a predistortion linearizer can do is produce an *ideal limiter* characteristic. Despite this limitation, it is possible for a linearizer to provide large benefits in signal quality when output power is reduced from saturation. Some improvement is possible even at saturation and beyond as the linearizer can correct for post-saturation phase distortion and power slump – but this improvement is usually very small.

Since the power out of the amplifier (in dB) is

$$P_{outA} = P_{inA} + GA = P_{outL} + GA = P_{inL} + GL + GA$$

Equations (13) and (14) can be rewritten referenced to the power into the linearizer (P_{inL}), and the desired transfer characteristics of the linearizer expressed as follows:

$$GL(P_{inL}) = GL_{ss} + GA_{ss} - GA(P_{inL} + GL(P_{inL})) \quad (15)$$

$$\Phi L(P_{inL}) = \Phi L_{ss} + \Phi A_{ss} - \Phi GA(P_{inL} + GL(P_{inL})) \quad (16)$$

Equations (15) and (16) can be solved iteratively for the ideal linearizer response needed to correct a given amplifiers transfer response. Figure 10 shows the response needed to correct a typical class A MESFET SSPA. As saturation is approached the rate of gain and phase change become infinite.

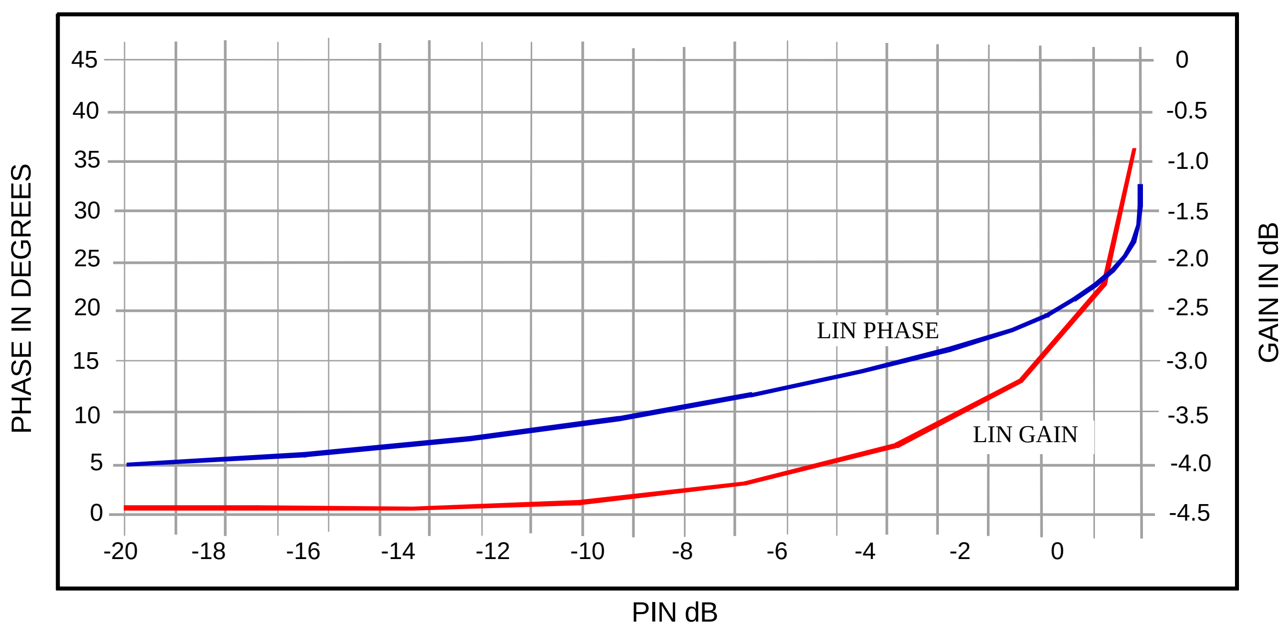

Figure 10. An ideal linearizer response requires the rate of gain and phase change to become infinite as SAT is approached.

$$dGL/dP_{in} = \infty \text{ and } d\Phi L/dP_{in} = \infty \text{ as } P_{out} \Rightarrow Sat$$

Such a characteristic can not be achieved in practice. Often a small amount of gain *expansion* near saturation, due to the finite dGL/dP_{in} available, is traded for superior C/I near saturation at the expenses of degraded C/I at higher OPBOs.

Where N is the number of carriers and Pav is the average power of the overall signal. For 4 carriers the OPBO for no IMD increases to 6 dB. The C/I for an ideal limiter driven by an infinite number of carriers (of random phase) is also show in Figure 11. The infinite carrier case is also known as noise power ratio or NPR.[1] Although the OPBO required for a given C/I increases with N, the improvement provided by PD linearization also increases with N.[12]

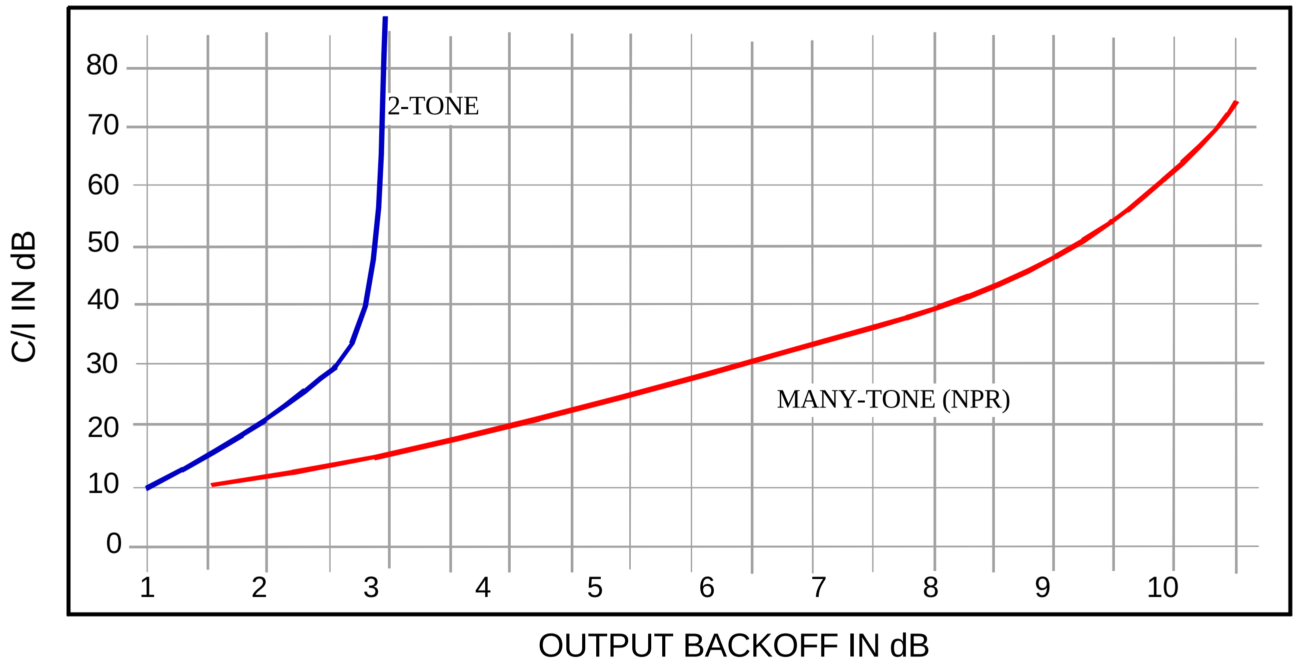

Figure 11. C/I of an ideal transfer characteristic for 2 and an infinite number of carriers.

Figure 11 shows the 2-tone C/I achievable by an ideal transfer characteristic. The C/I goes to infinity for OPBO greater than 3 dB. This result occurs because the peak-envelope-power (PEP) of a 2-tone signal is 3 dB greater than the average power. A signal backed-off by more than 3 dB never experiences clipping at saturation, and is subject to only a linear response. However to achieve this same level of performance with a larger number of carriers requires a greater level of OPBO. This is a consequence of the increase in PEP with carrier number:

$$PEP = N\,Pav \qquad (17)$$

Figure 12 shows the transfer characteristics of a SSPA and the corresponding corrected response resulting from linearization. The characteristics are for a class A power MESFET amplifier. Not how the shape of the linearized P_{out}/P_{in} curve approaches the desired ideal limiter characteristic of Figure 9. The separation of the 1 dB CP from SAT is a good indicator of linearizer performance. Ideally the 1 dB CP is located 1 dB in input power beyond SAT. Although the change in 1 dB CP in this case is not as great as for more non-linear amplifiers, the benefit can still be substantial.

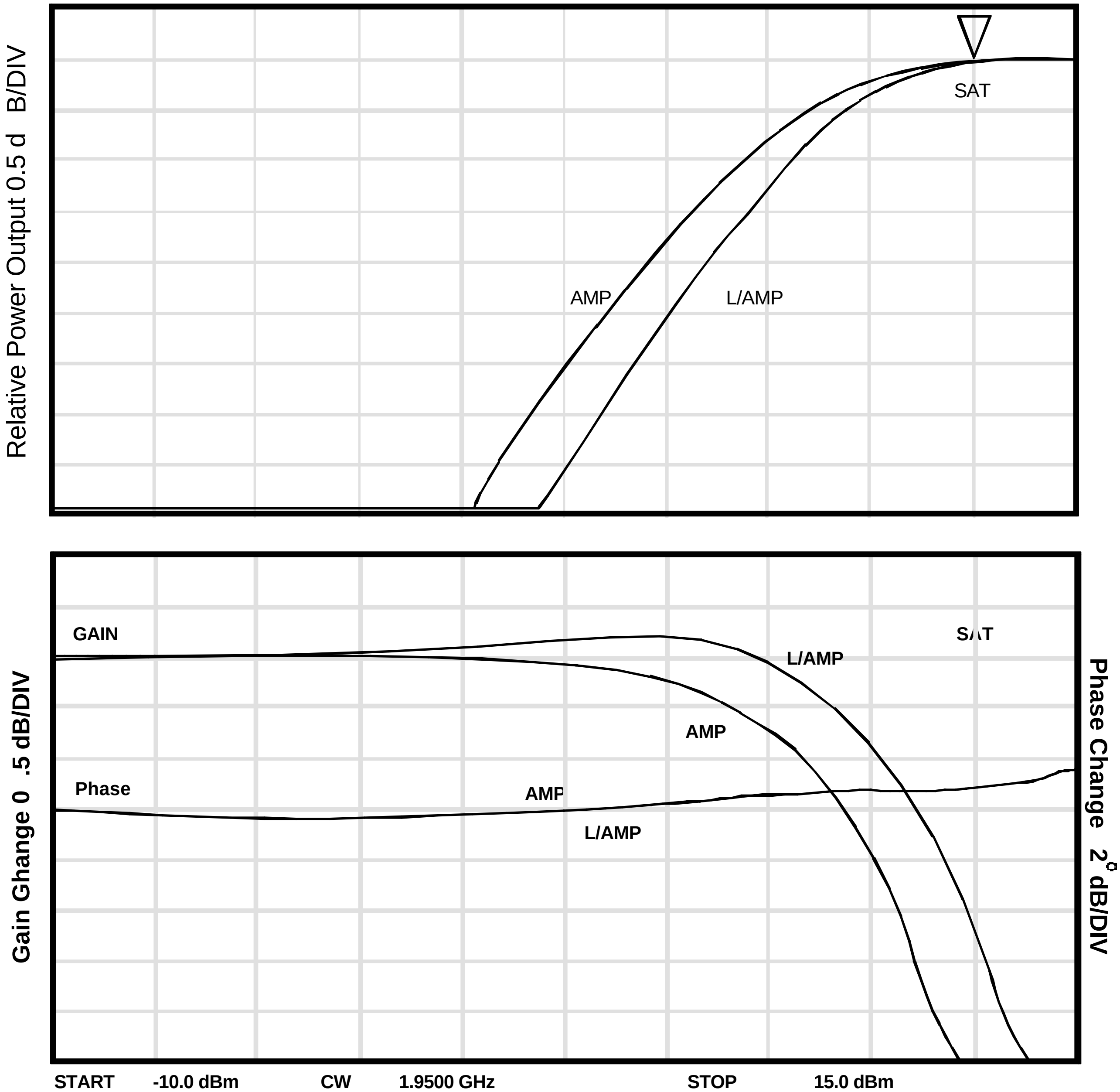

Figure 12. Transfer characteristics of a Class A MESFET SSPA and linearized SSPA.

Figure 13 shows similar graphs for a class AB amplifier. Here the 1dB CP has been moved nearer to saturations by almost 5 dB. The improvement in two-tone C/I as function of OPBO achieved by using a PD linearizer with class A and class AB MESFET SSPAs is depicted in Figures 14 and 15 respectively. The advantage of of linearizering SSPAs varies greatly with bias class and device type. The class A amplifier of Figure 14 shows only about a 0.5 dB increase in output power for a C/I of 26 dB, but a 2.5 dB power increase for a 50 dB C/I. The calsses AB SSPA, Figure 15, shows about a 1.5 dB increase in output power for a C/I of 26 dB. Ordinarily the more linear a SSPA, the less the advantage of linearization. When designing an SSPA to be linearized emphasis should be placed on optimizing parameters other than linearity.

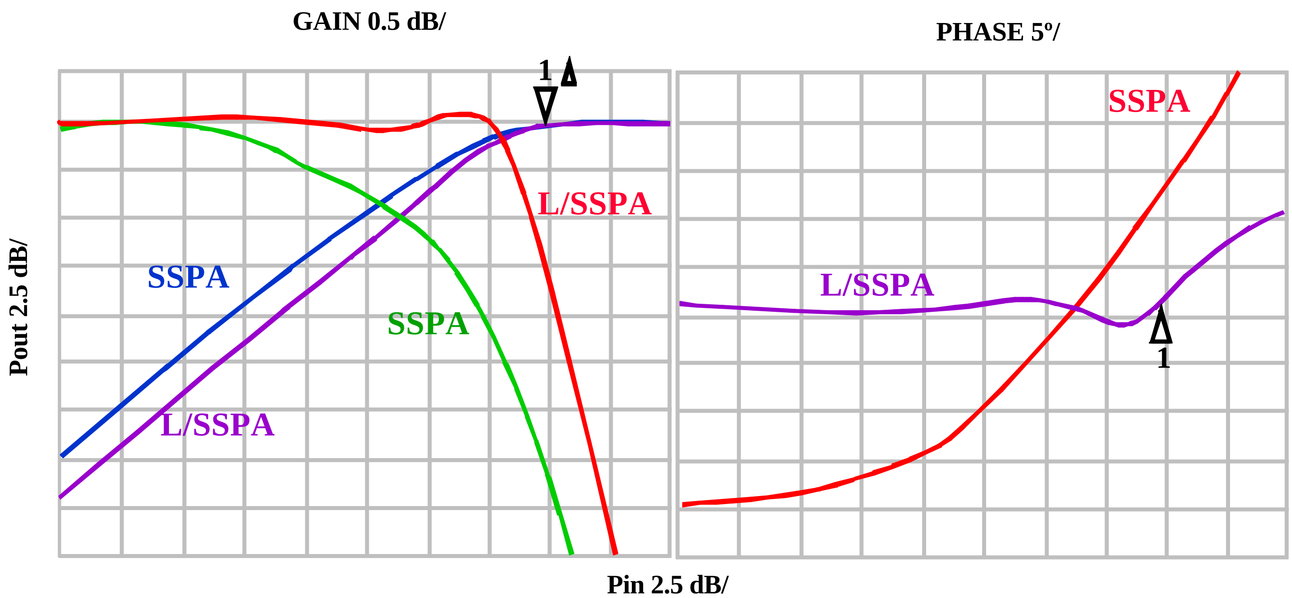

Figure 13. Transrfer characteristics of a Class AB MESFET SSPA and linearized SSPA.

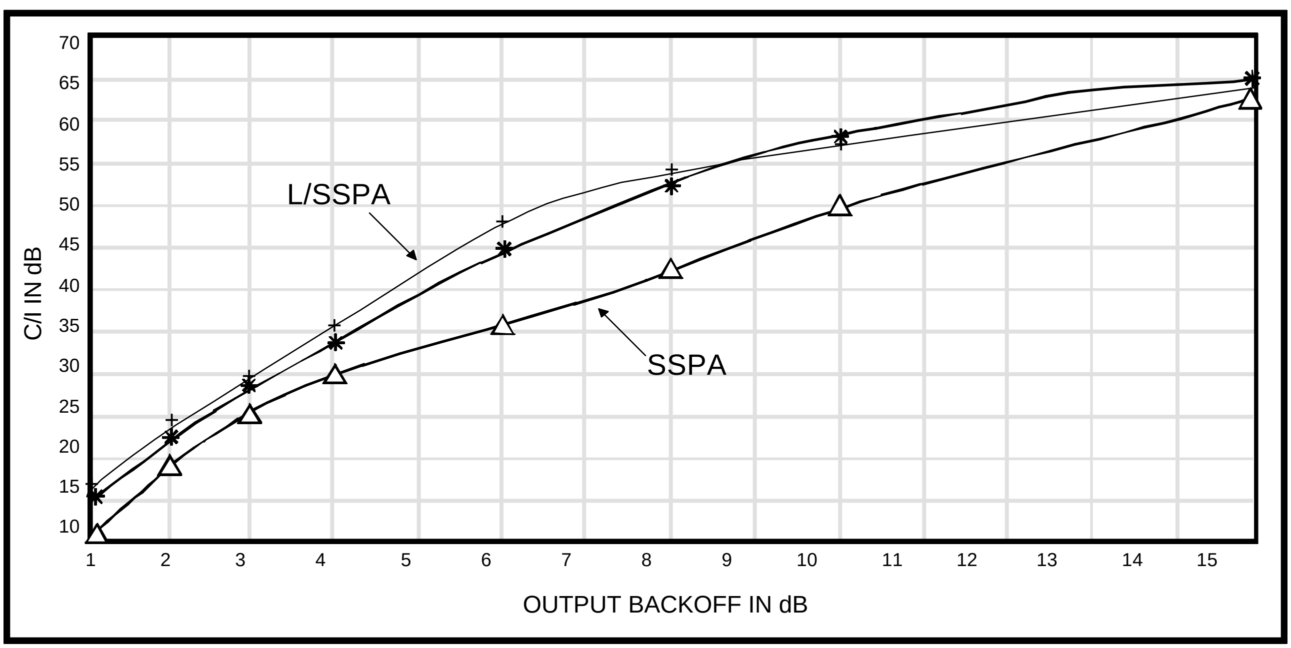

Figure 14. C/I vs. OPBO of a Class A SSPA and linearized SSPA.

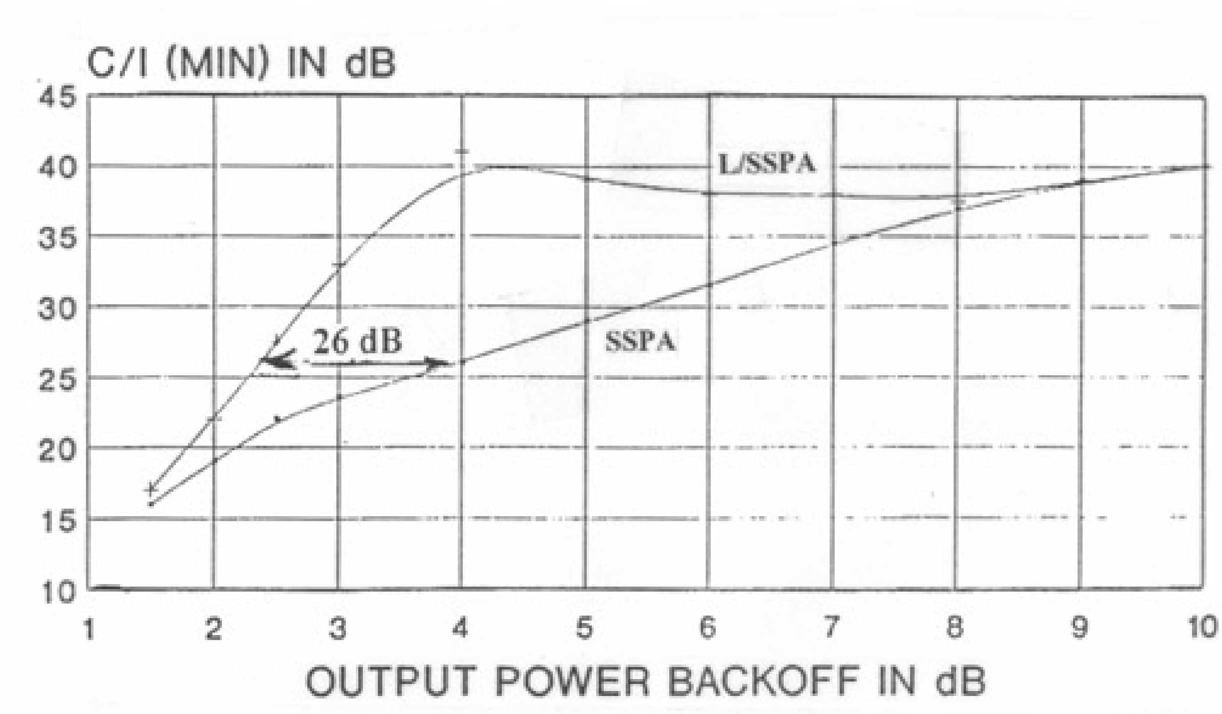

Figure 15. C/I vs. OPBO of a Class AB SSPA and linearized SSPA.

Design advances have greatly improved PD linearizers. Past linearizers were limited in bandwidth and difficult to adjust. New designs can offer greater than octave frequency performance, and the complex non-monotonic transfer responses needed by some SSPAs. They are much smaller in size and provide enhanced performance with easy alignment and excellent stability.

ADAPTIVE LINEARIZATION

For high linearity application (C/I > 50 dB) adjustment and maintenance of optimal linearizer settings become very critical. A change in phase of less than a degree can move a linearized amplifier out of specification. As a result of this parameter sensitivity, much effort has been devoted to the development of linearizers that can automatically adapt to environmental and stimulus changes. These adaptive linearizers can be considered a form of IFB linearization in which the feedback is applied to PD and FF linearizers. A measure of the linearizer's performance is generated. This performance measure (V_{PM}) can be the error voltage of equation (12), or a voltage proportional to the integrated IMD present in an unoccupied portion of spectrum near the desired signal. A microcomputer is normally used to analyze V_{PM} and determine the optimum linearizer settings.

In a FF linearizer, the microcomputer could be used control A0 and Φ0 in the signal loop and A1 and Φ1 in the cancellation loop – see respectively equations (7) and (8), and Figure 3. Using a search algorithm the computer would vary these parameters so as to keep V_{PM} at a minimum value.

In a PD linearizer the desired non-linearity can be produced using a power series:

$$Vout = k1\,Vin + k2\,Vin^2 + k3\,Vin^3 \quad (18)$$

Vin^2 and Vin^3 can be generated using double balanced mixers. Vin is applied to both ports of the mixer to obtain an output of Vin^2. A second mixer is used to obtain to Vin^3. If needed, additional mixers can be used to obtain even higher powers. The values of coefficients k1, k2 and k3 could be controlled by the microcomputer. As in the FF example, the computer would use a search algorithm to vary these coefficients so as to keep V_{PM} at a minimum value. Two non-linear PD elements can be combined in an arrangement similar to the Cartesian feedback system of Figure 7 to keep both gain and phase optimal.

The adaptive linearizers just described do not have the frequency response limitations of feedback linearizers, since they not attempt to correct for changes in the signal's envelope. These linearizers respond slowly to gradual changes in the systems characteristics.

TIME DEPENDENT AMPLIFIER NON-LINEARITY

The linearity of HPAs can degrade when operated with signals of bandwidths greater than several MHz. This degradation is referred to as *Dynamic* bandwidth and is a measure of the ability of the linearizer/amplifier to function with a rapidly changing signal envelope. It is particularly a problem for class AB SSPAs, and is not normally corrected by conventional linearization techniques. Figure 12 shows the change in 2-tone C/I as a function of carrier spacing, at 4 dB OPBO for a 40 watt, C-band, MESFET SSPA, designed for satellite service. (SSPAs used on satellites are normally operated class AB for greater efficiency.) The level of both upper and lower, third and fifth order IMD products are shown. The non-symmetry of upper and lower products is an indicator of the relative levels of AM/AM and AM/PM distortion.[10] As carrier frequency spacing is increased, C/I decreases noticeably. At a spacing of 30 MHz, C/I degrades by more than 2.5 dB. At 60 MHz, C/I is

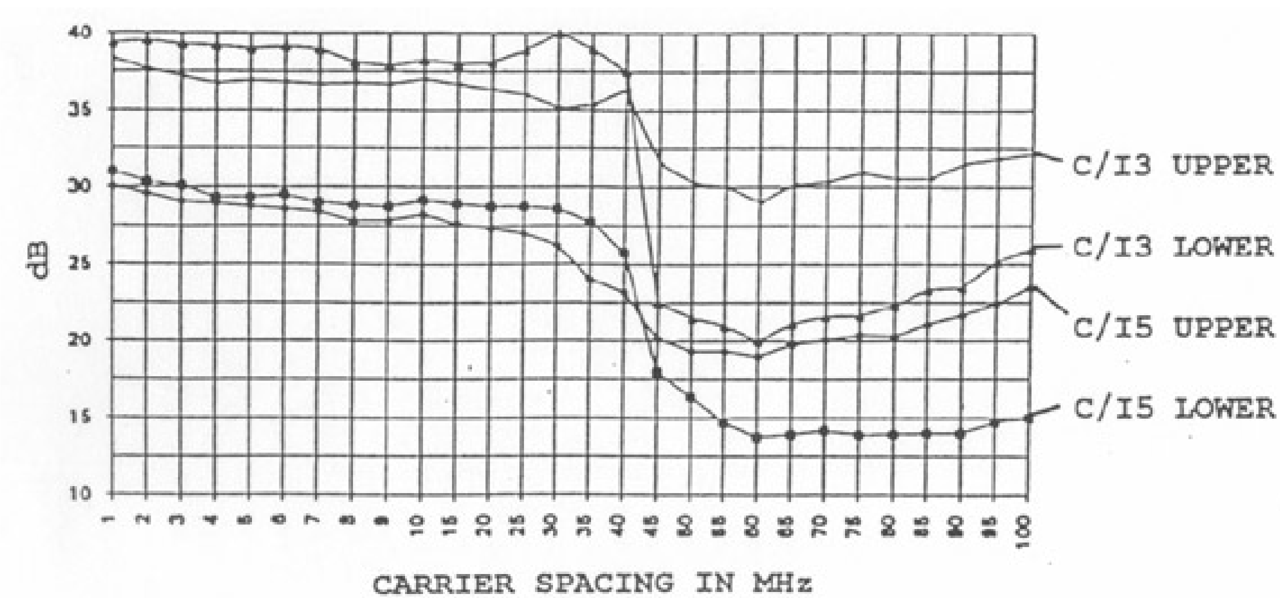

Figure 16. The linearity of SSPAs can degrade with increased carrier spacing

down by more than 6 dB. Some SSPAs have been found to degrade by more than 10 dB, for 100 MHz carrier spacing. NPR displays the same phenomenon. A similar effect is often observed for very close carrier spacing (< 50 kHz).

Cause of Degradation and Solutions

When the frequency separations of carriers is increased, the resulting signal's time-envelope changes more rapidly. For 2-carriers, the frequency of this envelope is

$$Fe = F\Delta /2 \tag{19}$$

where $F\Delta$ is the carrier separation. Distortion of the envelope produces IMD. The time-envelope, and related linearity, are dependent on the shape of the amplifier's gain and phase transfer characteristics. Assuming an amplifier with a flat frequency response, these parameters are normally considered time-invariant. When C/I changes with $F\Delta$, this indicates that the system is time-dependent. Time related change is particularly troubling for linearized amplifiers. In such systems, the amplifier transfer characteristics are corrected to provide a constant gain and phase, with input power level. Change with $F\Delta$ will cause the linearizer to be in alignment for one particular frequency, and out of alignment for another.

Figure 17 shows the 2-tone gain and phase transfer characteristics, as a function of carrier frequency separation, for the amplifier which produced the C/I performance of Figure 12.[13,14] The shape of the two-tone gain curves, (at Fc < 10 MHz), were in very close agreement with the amplifier's single carrier transfer response. The 1 dB compression point is degraded at higher separation frequencies. More significant is the dramatic change in phase transfer characteristics above 30 MHz; this is the frequency where the non-symmetry of upper and lower C/I products increased abruptly. This change is likely due to a resonance in the SSPAs bias circuitry.

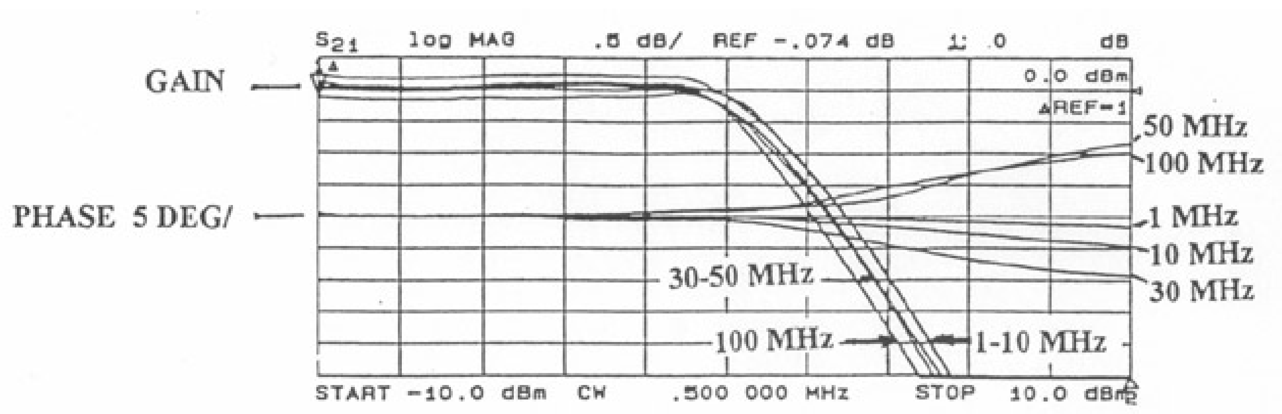

Figure 17. Two-Tone gain and phase transfer response of C-Band SSPA.

The principal cause of the SSPAs linearity degradation is believed due to changes in the bias currents. The uniform gain characteristics and the slump in saturated power at higher carrier spacing, indicates a problem in the drain circuitry. A significant induced voltage was found on the drain lines, when 2-carrier and NPR excitation was present. Induced voltages were also observed at the gate, but at a much lower level. Evaluation of the drain capacitors (7 µf) revealed they resonated near 0.5 MHz. Smaller low inductance capacitors were added on the device side of the drain circuitry. This change enhanced the 2-carrier transfer characteristics and improved amplifier linearity at wider carrier spacing. The resultant C/I as a function of OPBO for a 30 MHz carrier spacing is given in Figure 18. Performance of both the SSPA and a linearized SSPA, with and without the added capacitors is shown. For reference, the C/I with a 1 MHz carrier spacing is also included. Only one curve is shown, as performance with and without capacitors was essentially identical at 1 MHz.

SUMMARY

Linearizers increase SSPA power capacity and efficiency when high linearity is required for the transmission of digitally modulated signals and multi-carrier traffic. The greatest benefit is accrued with class B and AB amplifiers of marginal linearity and in applications requiring a very high linearity. In these cases linearizers can deliver a greater than 3 dB increase in power capacity, and more than double SSPA efficiency. Generally feedforward linearization is most valuable for applications requiring very high linearity. Indirect feedback also works well for these applications, but is limited to narrow bandwidth signals. Predistortion has the advantage of relative simplicity. It works over wideband widths and is viable for applications requiring both low and high linearity.

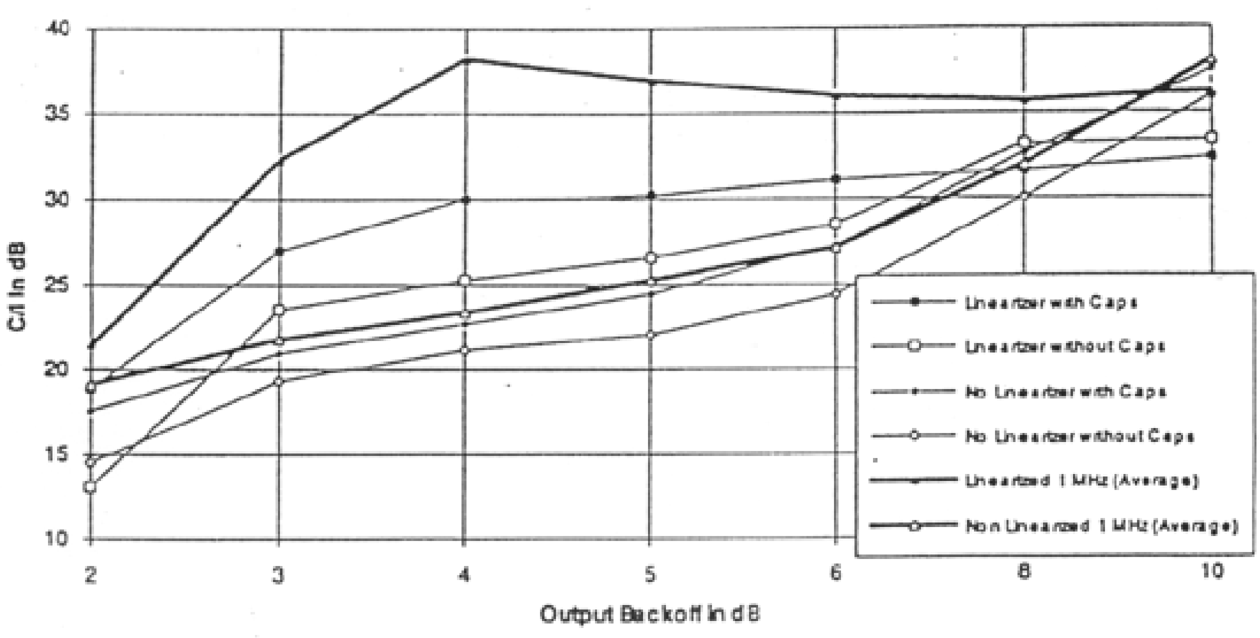

Figure 18. Improvement in C/I resulting from added drain capacitors.

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