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NOVEMBER 1984
LINEAR AND EFFICIENT BIPOLAR TRANSISTOR RF AMPLIFIERS

USING ENVELOPE FEEDBACK

BY

COLIN R. SMITHERS, B.Sc., LTCL.

A Thesis submitted for the degree of
Doctor of Philosophy
In the Department of Electronic and Electrical Engineering
University of Surrey

August 1985
After an introduction to amplifiers in communications and an exposition of the literature specifically relevant to high linearity power amplifiers, this study investigates more thoroughly various aspects of envelope feedback as applied to Bipolar Tuned Power Amplifiers at HF and VHF.

It is discovered that under the correct conditions a new mode of linear operation exists where gain compression, AM-PM conversion and input impedance are simultaneously linearised, and in this region DC-RF power efficiency is also improved. Spectral measurements are presented from an envelope feedback amplifier constructed to operate over this region.

A computerised system is described for measuring accurately the gain and phase shift of the test amplifier against variation of collector supply, quiescent bias current and RF drive power. The results from these measurements are presented as 3-dimensional projections and as contour plots. Subsequently the stored data is used to re-construct two-tone spectra, which is then analysed to show contributions to the spectrum from the gain compression and AM-PM conversion mechanisms separately. Conclusions are drawn with respect to effects of bias on these two mechanisms. A mechanism has been discovered which gives a symmetric spectra without requiring AM-PM conversion at the fundamental frequencies.

An attempt is made to model the amplifier with a non-linear circuit transient analysis program (SPICE). Good correlation is obtained for some parameters and these results are also plotted in 3-D and in contour.
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1. INTRODUCTION TO AMPLIFIERS IN COMMUNICATIONS

In general, electrical signal amplifiers should produce minimal distortion of the input waveform and should minimise electrical power wastage. This is particularly true in most aspects of communications engineering. It is also generally true that linearity and efficiency are mutually incompatible qualities.

Owing to the wide range of uses of amplifiers, there are many types and definitions of amplifiers. This study has been primarily concerned with those types used at the output stages of radio transmission systems and thus the definitions used are those common to this field. Even within this limitation, however, there are problems of standardisation of definitions. To establish an initial perspective, some basic amplifier characteristics, together with some established families of communications technology and corresponding fundamental problems, will now be outlined.

1.1 Signal Magnitude

Amplifiers may be classified according to signal magnitude i.e. large- or small signal amplifiers. Large signal amplifiers (specifically power amplifiers) are usually defined as having their maximum signal capacity limited by physical properties of the amplifier such as amplifying device saturation and/or the power supply rail voltage. This maximum may be specified in some cases as a distortion limit. Such amplifiers are often designed
Introduction to Amplifiers

such that some of the actual amplifying devices are operated into areas which are conventionally termed 'cut-off' regions. It should be noted that the term large signal amplifier does not imply a particular power (or physical size). On the other hand, small signal amplifiers are designed, ideally, to keep device characteristics constant over the whole signal magnitude range. They have their signal magnitude capacity limited by some more stringent distortion criterion. A small signal amplifier would never operate devices into cut-off or saturation regions. It will be clear that the distinction between these categories can become elusive in intermediate cases.

1.2 Amplifier Linearity

The term linearity nominally pertains to the input/output relationship of the amplifier:- Consider the input ($V_{in}$) to an amplifier as being a sum of sinusoidal voltages:

$$V_{in} = \sum_{i=1}^{n} V_i \sin(w_it)$$  \[1.1\]

A linear amplifier having an output/input voltage characteristic of the form:

$$V_o = kV_{in}$$  \[1.2\]

will produce an output containing only amplified versions of the input sinusoidal voltages.

A non-linear amplifier having a $V_o/V_{in}$ relationship of the
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\[ V_o = k_1 V_{in} + k_2 V_{in}^2 + \ldots \]  

[1.3]

will produce an output containing not only the original sinusoidal voltages (amplified) but in addition components at frequencies which are products of all the input frequencies. In the case of a single sinewave input, the output will contain the original plus direct multiples (called harmonics) and also a DC component:

\[ V_o = k_1 V_1 \sin(w_1 t) + k_2 (V_1 \sin(w_1 t))^2 + \ldots \]  

[1.4]

Frequency components arising as products of multiple frequency inputs are called InterModulation Products (IMP's) or InterModulation Distortion (IMD). (For this chapter only general aspects of IMD will be considered and not the precise nature of the non-linearities).

In the limit all amplifiers are non-linear. In fact, even the simplest passive components are non-linear\(^1\). Having said this, some RF power amplifiers are specifically described as 'linear'. As a contrary example, certain types of power amplifier (i.e. those used for the output stages of radio transmitters employing constant magnitude modulation methods such as Frequency Modulation (FM) and certain types of data modulation) require a low output of carrier frequency harmonics. This implies good shape conservation within one RF cycle of the input waveform. However, such amplifiers are not required to give close correlation between input and output signal amplitudes and, in
Introduction to Amplifiers

practiced, are often designed specifically to keep a constant output regardless of input amplitude. For the purposes of this study this type of amplifier is regarded as distinctly non-linear.

The parameter of RF envelope (peak-to-peak magnitude) gain constancy is critical to linear amplifier performance in a multiple frequency situation. Moreover, any waveform which has a time varying magnitude (e.g. a single carrier which is amplitude modulated) is a multiple frequency waveform and can be analysed both mathematically and physically to show 'sidebands' or extra frequencies located symmetrically at the sum and difference of the carrier and modulation frequencies. Indeed this was first shown practically in 1875 by Meyer in experiments with tuning forks and an acoustic modulator implemented by a rotating baffle with holes in it. This was proved theoretically nineteen years later by Lord Rayleigh. The practical result of non-linearity on an Amplitude Modulated (AM) waveform like this is the production of harmonics and IMP's. The harmonics and some of the IMPs may be removed by filtering, however the low odd-order IMPs (i.e. \(2f_1 - f_2\), etc, where \(f_1\) and \(f_2\) are single, uncorrelated carriers) lie adjacent to the wanted modulation frequencies and thus defy practical filtering methods.

Within communications engineering, different fraternities have different definitions of how linear an amplifier must be to earn this description. One measure of linearity is the relative output level of IMPs and (especially for wideband amplifiers) harmonics with respect to the wanted frequency components, possibly measured using a special test waveform. For example, a transmission combining amplifier feeding a common antenna at a mobile radio transmitting site (e.g. public services, Private Mobile Radio (PMR) etc) or alternatively on large naval vessels,
3rd order IMP's of 60-90 dB below wanted signals would be typical. On the other hand for a military Single Sideband (SSB) speech system, distortion limits are 40 dB in the adjacent communications channel, which in practice implies only 25dB 3rd order IMD performance.

Several standard linearity tests exist, individual tests being favoured within the different fraternities. Five are of major significance:

1) The Two-Tone Test

The use of two tones of equal amplitude spaced apart by a frequency difference unrelated to the two individual tones is the most common linearity test in radio engineering. It is particularly significant because the envelope traverses the entire power range of the amplifier (figure 1-1). It is most common to judge this test by the level of individual pairs of lower order products. In the accompanying spectrum the major distortion products are shown.

For amplifiers (and mixers) with weak non-linearities in systems where even the lowest IMD powers are troublesome, amplifiers are widely characterised in this test by the so called "intercept" method. The nature of weak non-linearities is that IMP amplitude increases with signal amplitude at a rate which is proportional to the order of the IMP, i.e. for a nth-order product, \( M_n \), the relationship is

\[
M_n \propto n f(t)
\]

where \( f(t) \) is the two tone waveform. This subsequently implies:
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a) That the signal-to-interference ratio decreases at the rate of \((n-1)\text{dB}\) per \(\text{dB}\) increase in input signal level

b) Eventually the third order component will be at the same level as the wanted component.

In practice this point is called the (third order) intercept point (figure 1-2) and it is generally specified by measurement of third order products. It is not a real point as there is a fundamental limit to the process at higher signal levels, and it must therefore be extrapolated from the more linear region of operation. It should be emphasised that the relationships generally hold well at low distortion levels, but may be very significantly in error at higher powers.

ii) 1-dB Compression Point

Especially owing to the failure of the intercept point technique at higher powers, it is common to define linearity in communications system components (e.g. general purpose RF amplifiers) by the amount of gain compression, i.e. reduction in gain with increasing input signal power level. 1dB is a common gain compression point at which the output power is specified.

iii) The Three-Tone Test

The spectrum of a colour TV transmission contains three major lines. These correspond to vision carrier, sound carrier and colour sub-carrier. It is not surprising therefore that a three tone test with the tones at the same relative levels is that used in TV engineering. This "Triple-Beat Test" is assessed for various products, but particularly by the level of one particular in-band product. (Figure 1-3)
Introduction to Amplifiers

**Amplitude**

**FREQUENCY**

**Time**

**FIGURE 1-1. THE TWO-TONE TEST IN TIME AND FREQUENCY**

**FIGURE 1-2. THIRD ORDER INTERCEPT**

**FIGURE 1-3. THE THREE-TONE TEST IN TIME AND FREQUENCY**

7
iv) Frequency Band Limited Noise

If measured in a narrow bandwidth and averaged over a long period of time, SSB has a spectrum which is not dissimilar to band limited noise. Thus band limited noise can be used to more quickly and consistently assess the likely effects of non-linearity on speech. The noise band may be amplitude weighted versus frequency within the band to more closely imitate speech. This is then called psophometric noise.

v) Noise Band Notch

In wideband FDM amplifiers where the individual signals are invariably un-correlated, the IMPs appear in effect as broadband noise. The usual technique of imitating the FDM signal is to generate broadband noise and to insert a frequency notch in this, subsequently measuring the IMD noise level in this notch.

1.3 Bandwidth

In particular applications the required operational bandwidth of the amplifier may be narrow or wide. A convenient definition of wide- (or broad) band is that of greater than one octave (i.e. greater than 2:1 frequency range between band edges, sometimes very much greater than 2:1) and hence harmonics of signals at the lower edge will fall effectively in-band, again rendering filtering powerless. This is not true of narrow band (or tuned) amplifiers and thus the whole design philosophy is different.

Again, the terms are user-dependent. For example, in the case of VHF PMR output amplifiers, the nationally allocated bands
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of operation are a few tens of percent wide. From the description above these amplifiers would be classed as narrowband. They are not, however, because the previous thermionic technology required individual tuning of amplifiers even within these bands. The transmitter typically consisted of a multiple switched-crystal oscillator followed by frequency multiplying amplifiers and concluding with an output frequency driver and PA. With so many cascaded tuned stages, broadband engineering was difficult and uneconomical. If a particular operator required multiple channels, these would be issued at relatively close spacings within the allocated band to accommodate the bandwidth limitations of the transmitter. It was only with the advent of frequency synthesizers that broadbanding has appeared.

Thus the bandwidth terms are relative and historically linked.

1.4 Class and Efficiency

Good power efficiency is always desired. This is because;

i) Power costs money (especially true in high power broadcasting).

ii) Energy storage involves cost, weight and volume, and must be replaced, which may cost time and resources (especially true in portable civil and/or military communications).

iii) Power wastage is generally in the form of heat, which often requires additional cooling to that provided by natural convection.

iv) High temperatures and thermal cycling in particular cause premature ageing and hence reduced reliability.
Introduction to Amplifiers

Accepting these facts there are two distinct approaches to amplifier design in which efficiency is the most affected parameter (considered from the point of view of wasted power). The first is to produce the output waveform by the approximately linear process (i.e. with no discontinuities) of developing a potential difference across a normally resistive load impedance by passing a current through this impedance, the current being controlled by the transfer relationship of some chosen active device (e.g. grid voltage-anode current in a valve or base current-collector current in a bipolar transistor). The second is to use the active device purely as a switch, deriving the required output waveform by other techniques. Depending on which other techniques are used, the amplifier as a unit may or may not be nominally "linear" as defined for this study.

In between these two approaches there are many widely used variations which can cause the type of an individual design to become indistinct. The term "class" is used to describe approximately the variation being used. However the class system (being a hangover from thermionic valve days when in practice the two methods were not so distinctly considered) has become extended and therefore somewhat confused. In conventional linear amplifiers (i.e. with at least one of the amplifying devices in a controlled linear state at any time) efficiency is normally controlled by the linearity constraints. This is accepting that linearity always becomes worse near the extremities of the operating characteristics, which is where efficiency is improving the most. However, as power is chiefly lost by the ohmic mechanism of active device dissipation, it is possible to save most of this power by switching the device(s) such that the device is either fully saturated or fully non-conducting for most of the time. This is known as "switched-mode".
Introduction to Amplifiers

In the early days of thermionic valves, operation was split into classes according to valve conduction angle i.e. that portion of one cycle over which device output current flows with the same waveform as the input waveform. Each class has an approximate efficiency associated with it.  

Class-A The normal, small signal amplifier is of class-A, conducts for the whole cycle and is not usually very efficient at all (dependent on required distortion performance). The maximum theoretical efficiency is 50%, which drops to 25% for a two-tone waveform. Zero signal conditions do not alter the power consumption in this class, efficiency then tends to zero.

Class-B Conduction for 180°. Used for tuned amplifiers (where the tuned output circuit extracts chiefly the fundamental output component of the output current) and also in anti-phase pairs for untuned amplifiers (where each device only conducts for one polarity of the waveform). Theoretical maximum efficiency 78.5%

Class-C Conduction for <180°, (typically 120). Used only for tuned amplifiers where the output waveform can be restored by a tuned circuit. Usually used in continuous amplitude applications or when amplitude modulation is applied to the final amplifier through power supply variation. Maximum efficiency dependent on conduction angle, for 120° this is 85%.

Class-AB Pseudo-linear mode which is class-A for small signals but large signals move the amplifier essentially into class-B. It draws a usable compromise between class-A linearity and class-B efficiency. In the case of valve amplifiers this class is also sub-split in classes;
Introduction to Amplifiers

$AB_1$ The input waveform does not drive valve into control grid current.

$AB_2$ Grid current is drawn at higher input powers with subsequent dynamic reduction of input impedance due to loading of the input tuned circuit.

This latter class allows greater peak power and efficiency to be achieved from a given device but the dynamic input impedance can cause non-linearity if the driving amplifier gain is output impedance dependent.

These are the major classes, however recent additions which are not so universally recognised include:

**Class-D** By recognising that power wastage in amplifying devices is chiefly due to output electrode dissipation during conduction, this can be reduced by minimising the voltage-current product, integrated over the whole RF cycle. In the limit the output waveforms should ideally be square. Originally this was achieved in thermionic amplifiers by the use of a third harmonic trap in series with the anode of the output stage, implemented with lumped circuitry or with shorted transmission lines. This gives considerable improvement with little added complexity. In solid-state amplifiers it became more normal to use square driving waveforms and hence the term "switched-mode" is used. However, more recently, class-D has come to mean the use of two active devices in a push-pull arrangement which essentially form a two-pole switch to generate the rectangular voltage waveform.

**Class-E** In this class due to the nature of solid-state devices, particularly with respect to internal capacitances, different approaches to the type of output coupling network yield improved
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efficiency even over class-D. Single-ended amplifiers are the norm and efficiencies in excess of 95% are reported.  

Class-F This is another term for the original Class-D using harmonic resonators. Other names include "bi-harmonic", "polyharmonic","multiresonator","Class-CD","single-ended Class-D"and "high efficiency Class-C".  

Class-G Audio amplifier technique using multiple power supply rails and output devices, each selected during the waveform to maximise efficiency.  

Class-H So called "variproportional" technique where power supply rail is instantaneously held just greater than necessary by a switched mode power modulator. The output device therefore develops very little potential difference. This might also be looked upon as a switched mode amplifier with an active linear output filter to remove switching components and restore high signal frequencies.  

Class-S If the mark-space ratio of a Class-D stage is altered, the average, or DC component of the output voltage alters also. Thus, with suitable filtering, a switching stage can be made to efficiently produce any waveform provided that it is switching at a sufficiently higher rate than the highest component in the signal waveform.  

1.5 Main Radio Engineering Categories: Commonly used amplifier technologies and associated linearity problems.  

1.5.1 Broadcasting  

Broadcasting embraces a wide variety of frequencies and
Introduction to Amplifiers

techniques, each with its own problems. Due to the very high powers involved, efficiency is of paramount importance to all operators for financial reasons alone, apart from any energy disposal considerations. Main frequency bands are:-

a) LF, MF and HF Radio using Amplitude Modulation (AM), the oldest and most traditional form of broadcasting. Propagation characteristics and the availability of simple reception techniques were the original reasons for this combination, major changes being subsequently resisted owing to market forces. Western Europe is now moving away from this in favour of VHF FM in order to:-

i) Increase signal quality,
ii) Combat spectrum crowding
iii) Avoid multipath fading (the second path being via the continuously changing ionosphere)
iv) To avoid the ever increasing ElectroMagnetic Compatibility (EMC) problems from thyristor power control, television switched-mode power supply and line-scan transient radiations and many other sources.

Despite this trend, the Third World shows an ever increasing demand (including spectrum demand) particularly for the ability to cover large areas of land (not necessarily home territory) with few transmitters. Also, in the USA MF remains popular especially for fade free car reception. Recently there has been a major campaign to introduce MF Stereo 12.

Owing to the modulation of amplitude, envelope distortion in the transmitter gives rise to adjacent channel IMD interference. Also as transmitting sites are considered to be anti-social in the current age of increasing rejection of visual environmental pollution, there is more than just the obvious financial pressure
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to have multiple transmitter sites. This gives rise to other problems; i) IMD production in the transmitter output stages due to reverse injection of unwanted transmissions and ii) IMD and harmonic production by currents induced into metal structures due to electro-chemical non-linearities in mechanical couplings (the so-called Rusty Bolt Effect).

b) VHF FM Radio. Again, multi-user sites and common antennas give rise to IMPs from transmitter outputs, combining networks, and in non-linearities in the antenna itself. The transmitters have fewer problems with IMD because of the steady envelope nature of FM.

c) VHF and UHF AM Television. TV is generally broadcast as AM. This is for bandwidth considerations. Because of the very high powers involved in TV signals and small coverage areas (due to the propagation characteristics at the frequencies used and the large signal strengths required to achieve acceptable signal-to-noise performance in the 8 MHz wide channel), the energy cost is very significant to operators. If Television were to be broadcast as FM, the power cost would appear at first sight to be worsened. However, the nature of the composite vision and synchronisation signal is that there is a relatively high dynamic range required between the line synchronising pulse and the general vision level. This means that for good linearity the average efficiency of a TV transmitter is very low, 15% is not untypical (but it is picture dependent). FM transmission of TV pictures would be at higher efficiency, and at broadcast picture quality levels would enjoy the FM "capture effect" and associated signal to noise ratio improvements of wideband FM over AM. Indeed, satellite TV links usually are FM.

Due to i) the ability of the eye to detect non-correlated
Introduction to Amplifiers

dynamic interference and ii) the total inability of AM receivers to reject unwanted signals, there have to be strict adjacent channel interference limits which impose linearity, and therefore efficiency constraints.

Antenna engineering for TV transmission needs to be of a high standard. Antennas are said to be "matched" if they present the same impedance as the characteristic impedance of the transmission cable. If they are matched, all the cable power will be radiated by the antenna, otherwise some proportion of the power will be reflected back along the cable to the transmitter. The degree of match is commonly measured by "return loss":

\[
\text{Return loss (dB)} = 10 \log_{10} \frac{\text{Power Reflected}}{\text{Power Applied}}
\]

or by Standing Wave Ratio:

\[
\text{SWR} = \frac{(1 + S)}{(1 - S)}, \quad S = \text{Voltage Reflection Coefficient.}
\]

The effect to the transmitter of antenna mismatch is to present a load impedance which is not only incorrect but which varies across the modulation bandwidth. This arises because the currently transmitted signal is corrupted by a delayed version of the signal resulting from reflection of a portion of the original at the antenna mismatch. The transmission delay along the cable can be quite considerable with typical TV antenna masts. These reflections can not only add unwanted frequency components but also introduce a regular temporal- and hence spatial distortion in the picture which is singularly easy for the eye to detect. This is commonly called ghosting. The industry standard for antenna matching is an SWR of 1.05:1, which is a return loss of 30dB.
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1.5.2 Communications

a) HF Communications

HF is used for middle (greater than 30 miles) and long distance civil and military point-to-point and mobile applications only when other techniques are unsuitable. Continual international demand for spectrum allocations perpetuates this situation. Single Sideband (SSB) is the dominant mode for speech communications and a wide range of proprietary techniques are used for data, some of which have time-varying envelopes and some of which do not. There is currently a trend towards organising all data to be transmittable through a conventional SSB speech link. Single- or dual-channel (Independent Sideband, or ISB) used to be the most common form of transmission, however especially for military base stations (including ships) the trend is towards multi-channel systems based on multiple low-power SSB sources and one wideband amplifier/antenna combination.\(^\text{14}\)

These wideband systems are of very low efficiency due to the high peak to average ratio of a multi-frequency waveform and the consequent need for Class-A techniques to achieve very strict distortion criteria. The governing criterion is that for multiple signals being transmitted, there should be no mutual interference to any of the much greater number of signals probably being received simultaneously at the same, or at an adjacent site.

Typical mobile and portable SSB transmitter problems are harmonic radiation and adjacent channel interference due to odd-order IMPs. The harmonics are, by definition, at least one octave higher in frequency and can be dealt with by filtering. Efficiency, while being relatively high (typically 35% output amplifier efficiency for speech type waveforms), still allows
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room for 3:1 improvement in energy storage requirement. In practical systems this is normally defined to allow some minimum period at a chosen transmit:receive ratio, typically ten hours at 10:1. Localised heating and associated drift/reliability effects are also of major importance.

b) VHF and UHF Communications

Military, Public services (police, fire, ambulance, water, electricity and gas) Civil Defence networks, Private Mobile Radio [PMR] (business radio such as taxi or business and domestic field service engineering) and cellular radiotelephone networks all use VHF and UHF for point-to-point and mobile purposes. In Europe AM and FM speech are the dominant modulation modes. At various times in recent years SSB as been put under close technical scrutiny as an alternative on the basis of superior spectral usage and service area:power properties. Some technical difficulties exist that are peculiar to the use of SSB modulation particularly for mobile operation, but many of these problems have been solved to a large extent over the last decade. Apart from linearity these include fast fading which cannot be removed by conventional feedback AGC, frequency stability, Doppler shift and multipath reflections. Commercial and regulatory politics have also been prevalent in the arguments, though these are less visible in the literature.

As there are many and increasing numbers of mobile radio users in a normal town and due to the social pressure towards co-siting, there are ever increasing problems with extending combiner technology, both active (with amplifiers) and passive (with filters) and with the Rusty Bolt Effect.
c) Microwaves and Satellites

There is an apparently insatiable demand for worldwide telephone and data communications. Apart from extensive cable networks, the major traffic routes for this are microwave point-to-point and satellite links. Individual links are wideband, multiple speech and data signal being encoded both by frequency stacking called Frequency Division Multiplex (FDM) or by time sharing, usually of digitised information, called Time Division Multiplex (TDM). Both FDM and TDM have non-constant envelopes. In general, completely different amplifiers are used from those used at lower frequencies, travelling wave tubes being the dominant technology. Sometimes the amplitude linearity of the microwave link itself is not so relevant since the information band is frequency modulated onto the link carrier. The linearity problem then lies principally in the frequency modulator and de-modulator stages.

Television and radio broadcasting companies also use these types of wideband techniques in order to route signals both nationally and internationally. It is common practice to route TV as FM and Radio as Pulse Coded Modulation (PCM).
2. REVIEW OF NARROWBAND DISTORTION TECHNIQUES

This chapter reviews the history of narrowband amplifiers and linearisation techniques particularly as intended for use with single channel SSB transmitters.

2.1 Early History of SSB

SSB was first considered for use as a transmission method by Carson in 1915, after mathematical analysis of Amplitude Modulation (AM) revealed the presence of discrete sidebands. This appears to be independent of the acoustics experiments by Meyer in 1875 and the analysis by Rayleigh in 1894. Initially SSB generation for radio was achieved by using the tuned antenna system (at LF) as the filter.

In their attempts to wire the American nation with telephone during the Twenties, Bell Labs had problems with limited spectrum and took to using SSB for this. At this time the phenomena associated with imperfect amplifiers were all too well known and tended to be the limiting factor in FDM repeater systems, wider bandwidths incurring greater losses and therefore requiring more repeaters which in turn meant more distortion products at increasing power levels. (It had been observed that while the second harmonic distortion power rises with the root of the number of repeaters, the third order IMP power rises with each new amplifier.)

SSB continued to develop and be used on radio links, achieving some bandwidth reduction despite the lack of any systems for IMP cancellation. This implies that realistic levels
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of IMPs had to be universally accepted and amplifiers simply operated only over a sufficiently linear portion of their characteristics to achieve this performance, despite the lack of efficiency and the capital cost of the high peak power equipment.

During the second World War SSB radio was used to keep the States in contact with its armed forces in far parts of the world and in subsequent years widespread use was made of Independant SideBand (ISB) modulation in which both upper and lower sidebands are radiated. This gives two transmitted channels with a lower capital cost than by using two separate transmitters. Because of distortion problems in the adjacent channel however, it was not uncommon to operate the two sidebands at some frequency spacing away from the common carrier.

In the early Fifties the Federal Communications Commission (FCC) of the USA proposed rules requiring all HF point-to-point communications to use SSB. This gave rise to increased effort towards SSB technology generally.

In 1952 Kahn proposed a system which was aimed at enabling conversion of high power AM transmitters to SSB with minimum capital expenditure to the user and also giving the added benefit that the maximum power efficiency was in excess of that for conventional SSB transmitters, thus yielding great savings by avoiding the purchase of entire systems plus keeping running costs to a minimum. This was achieved by taking a high quality SSB signal and splitting it into two channels. One started with a limiter and then used the conventional class-C stages right up to the antenna while the other took out the original amplitude envelope information and passed it through a more or less conventional AM modulator to the final stage (fig.2-1). The system was experimentally applied to a 2 kW AM
Narrowband Systems

transmitter and resulted in a 2.5 kW Peak Envelope Power (PEP) output.

![Block diagram of narrowband systems](image)

**FIG 2-1 THE KAHN EXPERIMENTAL CARRIER ELIMINATION AND RESTORATION SYSTEM (from ref. 18)**

This system showed an interesting bonus over the conventional linear amplifier technique as its 3rd order IMP level was at -30dB with respect to either of the tones in a two tone test transmission (recognised as a severe test of an SSB system as its amplitude passes through all points from peak envelope power (PEP) to zero amplitude) and this was at least typical of normal transmitters but at much higher efficiency. It was predicted also that the performance could be made to reach 40dB with careful design. However, it was indicated that the cancellation of higher order products was very dependent on modulator bandwidth as the spectrum of the envelope signal (being in effect a rectified sinewave for the two tone case) contains
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components at frequencies far in excess of the difference frequency of the two tones.

It should be noticed that an 'output level control' is used, which appears from the text to be a form of Automatic Limiting Control (ALC) which not only limits the the maximum level but also raises the level of quiet signals, keeping them near the system maximum. The idea is also suggested of using ALC in a different mode of allowing carrier to be sent during quiet periods, thus aiding receiver implemented Automatic Frequency Control (AFC), Automatic Gain Control (AGC) and even squelch functions (muting the receiver during no transmission). The phase correction stage in the amplitude limited path is for equalising the time delay along the two paths. By 1956 Kahn had used the system commercially up to 400 kW PEP and with results occasionally reaching 40 dB IMP performance.

2.2 Early Distortion Reduction Techniques.

Also in 1956 W.B.Bruene produced a paper on "Distortion reducing means for SSB transmitters". Here direct RF feedback and envelope feedback were discussed and compared in reasonable detail.

2.2.1 RF Feedback.

RF feedback was claimed as a "very effective means of reducing distortion, 10 dB of feedback producing nearly 10 dB of distortion reduction". Two particular tuned amplifier circuits were described with some emphasis on the phase-gain performance of the various stages and their relationship with total amplifier stability. In conclusion it was stated that for a two stage (servo) tuned amplifier, 18dB of feedback could be used although
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it had been found in practice that limiting feedback to 12 or 15 dB gave a more readily engineered performance owing to the consequently wider stability margin.21,22

2.2.2 Envelope Feedback

The envelope feedback technique was carried over from AM practice, first mentioned by Terman and Buss in 1941. Their original figure is reproduced in figure 2-2. Breune modified this for SSB use by the addition of an AF amplifier, the gain of

(a)

(b)

Fig. 5

(a) Schematic diagram showing method of applying balanced feedback to a linear radio-frequency amplifier.

(b) Practical circuit used in tests.

FIG 2-2 TERNMAN'S ORIGINAL ENVELOPE FEEDBACK SYSTEM (From 23)
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which was dependent on input power level, this being necessary to maintain the same proportion of improvement at the lower power levels encountered in SSB signals with no carrier present (figure 2-3).

The important aspect of the way this feedback is applied is that the distortion from each of the envelope detectors cancels out because they are operated at exactly the same power level and only on the appearance of a difference (error) signal does any feedback correction take place.

Bandwidth requirements for the feedback path were again recognised as being large in comparison with signal bandwidth (not forgetting tuned circuit bandwidths also), and that choosing a frequency to equalise the feedback amplifier up to, effectively sets the system performance limits.

Bruene concluded his paper with a performance description of his three stage amplifier with RF feedback and also a form of

FIG 2-3 BRUENE’S CIRCUIT FOR ENVELOPE FEEDBACK

APPLIED TO SSB (From ref. 20)
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envelope modulation predistortion, obtaining a controlling waveform from a "modulating signal synthesiser" which is fed to the first control grid (fig.2-4). With this system he obtained 50dB IMP performance with a two-tone test signal.

Other important conclusions to be drawn from Bruene's observations are:-

1) Envelope feedback can easily provide a significant improvement in low order IMPs but it is difficult to obtain much improvement in high order products.

![Diagram](image)

**FIGURE 2-4. BRUENE'S BIAS BENDER CIRCUIT (From ref. 20)**

2) Direct feedback, when successfully applied, affects all products (provided that tank circuit bandwidths are sufficiently wide compared with signal modulation bandwidth).

3) Two-tone test measurements and noise band test measurements on the same amplifier can give different results (owing to different rates of decline of higher order products)

In 1965 Wood took out a patent on a TV transmitting
Narrowband Systems

frequency converter which employed envelope feedback around a varactor diode mixer (fig.2-5). This envelope technique was then applied by him to microwave amplifiers. How this differed from Terman's original article is not clear (apart from in the transition from thermionic to solid state devices) but nevertheless the publication does serve as useful extra publicity for this technique.

In 1979, Mendillo repeated Bruene's Terman/Buss development but with solid state devices and using some more elaborate synchronous detectors instead of envelope detectors. His results show the type of improvement that Bruene predicted, i.e. considerable improvement for the lower order products, this improvement reducing for much higher order products. At medium power levels the improvement is quite dramatic i.e. 30dB improvement in 3rd order IMPs at a PEP output of +22dBm, showing how crossover distortion can be successfully combated by this technique.

2.2.3 Bias Pumping.

There is a paper by Pontius of the E.F.Johnson Co. in which two conventional class-C 3-stage amplifiers were modified to class-AB in order to produce linear amplifiers of 12.5W and 30W PEP. It was stated that:

"Bias techniques were applied in such a way as to utilize the inherent characteristics of the devices to provide more linear performance."

Quite what these techniques were is not stated but they were presumably a shifting of the bias point with the incoming signal level. This would have the effects of:
Figure 1.

Figure 2-5. Wood's Mixer with Envelope Feedback (From ref. 24).
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a) Affecting the gain of the stage by a few percent, probably enough to correct for peak limiting for a certain range.

b) Affecting the AM-to-PM conversion characteristics of the stage by some amount.

Whether the system was a feedback process or a predictive technique is also unknown. The results are singularly impressive anyway, if only because of the way that the higher order products fall away rapidly. In the published spectrographs 3rd and 5th order products are at -40dB and -56dB respectively from either tone in the 12.5W module and at -35dB and -60dB in the 30W module.

Listening tests were performed with what is at first sight a peculiar arrangement (fig.2-6). It must be presumed that the filter present in the path from the interfering transmitter is tuned so as only to improve the receive selectivity on the wanted channel.

2.3 Recent Feedback Developments.

In 1979, after some years working on various high efficiency /high spectral purity systems (principally at Bath University), Petrovic and Gosling came up with a developed form of the Kahn transmitter; Polar Loop®, (fig.2-7). This was further developed at Bath into the Cartesian Loop system.

The Polar Loop system differs from the Kahn system by the use of:

1) A phase-locked mixer loop, enabling any transmitter frequency to be easily selected and any spurious phase modulation
in the system to be reduced by feedback (especially including any unwanted AM-to-PM conversion in the modulated power amplifier. This phenomenon will be covered more fully later).

FIGURE 2-7. THE "POLAR-LOOP" TRANSMITTER SYSTEM (From ref.31)
2) The use of synchronous rectification to obtain more accurate envelope signals.

When the published results are first examined the impression is of a good system;– the 3rd order IMPs are well attenuated at -50dB. However practical systems and closer analysis show again that judging adjacent channel transmitter performance by 3rd order components alone can be misleading in terms of actual system performance:– In a typical speech waveform it is the higher order IMP's which fall in the 1st and 2nd adjacent channels and this (as shown by Bruene) when dealing with a single quadrant signal processing technique (i.e. when the feedback signal is a function of the magnitude of the RF signal and not its instantaneous phase polarity) requires a much wider bandwidth in the envelope feedback channel owing to the rectified nature of the feedback signal. Also, consider a qualitative analysis of the phase-locked-loop under two-tone conditions (fig.2-8):– The vector sum of the two tones can also be treated as a single

![Diagram of vector addition of two tones](image)

Phase of resultant: 0, 0, 0, 180, 0, 0, 180

**Figure 2-8. Two-tone Analysis**
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frequency of varying amplitude which has a phase reversal each time the envelope passes through zero. This has to be copied identically by the PLL in the Polar Loop transmitter. In order for an oscillator with voltage controlled frequency to perform a $180^\circ$ phase shift requires an instantaneous departure to DC for a period of half of one cycle and then an instantaneous return. The bandwidth and transient response requirements are quite obviously very large and not necessarily related to the difference frequency of the two tones.

Although the two-tone test is a harsh and, debatably, an unrealistic form of test for an SSB transmitter system, it does reveal one reason for the disappointing practical results found by some engineers which were only a few dBs better than a conventional transmitter yet involving considerably greater engineering complexity.

The more recent Cartesian Loop system (also developed at Bath) avoids the problems outlined above. Figure 2-9 shows a block diagram for the system as implemented in a Third Method$^{31,32}$ type SSB transmitter. In terms of added complexity over the basic transmitter, implementation is clearly simpler than for Polar Loop and the performance is superior, published results claiming 70dB IMPs. By rectifying synchronously using already present reference frequencies ($w_{rf}$ and $w_{rf}+\theta$) the IMD is converted to baseband where it can be applied as negative feedback directly before the original frequency conversion. The quadrature nature of the system effectively specifies at any instant a particular output vector and any deviation from this, either in amplitude or phase, will thus be corrected. The specific improvement is that the feedback spectral density is directly related to that of the original modulating waveform, unlike Polar Loop.
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FIGURE 2-9. CARTESIAN LOOP SYSTEM (From 31)
3. REVIEW OF WIDEBAND DISTORTION TECHNIQUES

3.1 Introduction

In HF communications the need to re-tune a transmission quickly, which has resulted in the current trend towards broadband operation, is because:-

1) Changes in propagation conditions require frequency changes which interrupt radio links during the tune up period.

2) Communication systems architectures are currently and increasingly being based around banks of paralleled wideband power amplifiers for flexibility and reliability reasons.

3) Anti-intercept modulation modes such as frequency hopping and other more subtle spread-spectrum techniques require such high retune rates that, due to the mechanical engineering limitations of tuned transmitters being several orders of magnitude too slow, broadband techniques are imperative.

4) As systems become more complex so the necessary operator education level rises, therefore any moves toward reducing operator skill level requirements are welcomed. Broadbanding achieves this.

Wideband amplifiers have for many years concentrated mainly on achieving linearity of the active device itself and by ensuring that the amplifier is not overdriven or, if necessary,
Wideband Systems

by greatly overrating the amplifier. Circuit conditions that have been known to be important in valve amplifiers are:

A) Anode load impedance; i.e. not too high, so as to avoid large anode voltage swings.

B) Power supply regulation, especially with regard to bias supply impedance at envelope frequencies.

C) Resistive swamping out of grid non-linearities.

D) Adequate driver power capability (preferably of several times the required drive power).

E) Unwanted RF feedback to be avoided at all cost i.e. by careful attention to decoupling, earth loops, electrostatic (capacitive) and magnetic (inductive) screening etc.

These are also applicable to modern solid state amplifiers although possibly in a different order of priority owing to the very low impedance of these devices (typical input impedances are ~1 ohm). Point E should not be underestimated; screening cans frequently affect intermodulation and harmonic performance, when they are flexed or have their screws tightened the IMP's can be observed to vary.

3.2 Bandwidth Limitations; Distributed Amplification

Wideband amplifiers have a problem of their own as well in that the gain tends to drop off at high frequencies due to the effect of circuit plus device output capacitance in parallel with
Wideband Systems

the output load impedance. This gives rise to the well known gain-bandwidth product limitation. Paralleling devices does not help and the only technique to overcome this is the distributed amplifier,\(^3\),\(^4\),\(^3\),\(^5\),\(^6\) (fig.3-1). Here the amplifying device input and output capacitances form the capacitors of a lumped component artificial transmission line. It is by using this technique that the very successful 1kW broadband amplifier designed by Stokes of Marconi has remained unchallenged for 25 years and forms the basis of most modern broadband systems.\(^3\) Distributed techniques are also being used experimentally to improve and VHF and UHF solid-state device performances.

![Diagram of distributed amplifier principle](image)

**FIGURE 3-1. THE DISTRIBUTED AMPLIFIER PRINCIPLE**

3.3 Broadband Systems

The first broadband system,\(^3\)\(^8\) the Integrated Communications System ICS1 was conceived in 1957 and had the following important
Wideband Systems

transmission features:-

1) SSB operation.

2) Three semi-broadband antennas which together cover the whole HF band.

3) Drive units and broadband 1kW linear amplifiers which could be coupled to any one of the antennas.

4) Multi-couplers which could allow up to eight 1kW transmitters to be coupled to any one antenna.

ICS1 was replaced in 1965 with ICS3 which had some new features:-

1) A two-port wideband antenna

2) A wideband amplifier bank consisting of several of the ICS1-type amplifiers coupled together with hybrids.

3) Broadband active receiving antennas.

The Marconi amplifier is facing competition now from solid state amplifiers and in a comparatively recent report by Trollope and Lauder of ASWE compared IMP figures of -49dB, -64dB and -67dB for 3rd, 5th, and 7th order products respectively for the Marconi amplifier against -55dB, -71dB and -77dB for a "State-of-The-Art" (SOA) solid state amplifier that they had built for them. It is curious to note that no mention is made of harmonic performance.

The Naval Research Laboratories in the States proposed a
Wideband Systems

Wideband HF system architecture in 1980 which appears to be based on the fact of increasing reliance of HF in tactical communications due to recent revelations on satellite system vulnerability. There is also very considerable emphasis placed on Electromagnetic Compatibility (EMC) and control of Electromagnetic Interference (EMI). To this end there is numerical discussion of the intermodulation performance of transmitting and receiving systems and also predictions of the requirements of these elements in the context of a co-sited multi-user system which can operate simultaneously on all frequencies other than those actually in use elsewhere in the system. The paper also mentions that "the employment of negative feedback and use of large numbers of the most linear solid-state devices are expected to yield a -55dB performance" (which presumably harks on the ASWE work), but adds; "a modest improvement with respect to ICS3." Mention is made of Adaptive Interference Cancellation (AIC) work which is currently under investigation at the NRL.

3.4 External Linearity Limitations

There is some question as to realistic figures of system element requirements i.e. simply improving transmitter amplifier and receiver performance to ideal proportions ignores a more subtle limitation viz. the so called Rusty Bolt Effect. This is the situation where, for example, a ship's superstructure acts as parasitic element and the currents flowing in the superstructure also pass through the structural joins which appear electrically as many diode-like functions. These non-linearities in turn generate distortion products which are radiated by the superstructure and are obviously of a totally unpredictable nature.
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[Some work is being carried out currently at York University\textsuperscript{41,42} investigating the feasibility of the deliberate use of Rusty Bolt as an aid to corrosion location and progress monitoring in land-based civil engineering applications. Novel use is made of multiple time-encoded signals which reveal the fault location directly using techniques not disimilar to established commercial hyperbolic radio navigational systems, e.g. Loran.]

Quantitative information on the typical level of products to be found in practical situations is rare and very varied; ASWE quote \textsuperscript{39} the results of an earlier paper by Mason published in 1963 \textsuperscript{43} and claims that 60dB for 3rd order IMPs and dropping to 80dB for 5ths is a realistic figure. Examination of Mason's results \textsuperscript{43} reveals that this is a worst case view and that in fact 70dB for 3rds is more common, although it must be admitted that Mason did point out a typical daily measurement variation of 10dB which was thought to be related to weather conditions. The ship was of all-steel construction and the location for the experiments was anchored in open water away from other shipping and any shore environment. The NRL, on the other hand, discuss the findings of another NRL report \textsuperscript{44} where IMP measurements were made on three US Navy ships that were "designed with deliberate attention to EMI control by the use of non-metallic topside fixtures where feasible, and careful attention to bonding of metallic structures." The results of this report and the conclusions subsequently drawn by Davis et al \textsuperscript{40} suggest that a figure of 110dB for 3rd order IMPs is more typical and that therefore 120dB is a necessary performance target for transmitter system performance.

The difference of opinion between ASWE and NRL emphasises the lack of coherent information on this subject. In the opinion
Wideband Systems

of this author, it will probably be found difficult to reliably achieve the NRL performance targets on even the simplest transmission interfaces at these power levels. Incidentally, the NRL report was referenced in the more recent ASWE report.

One other calculation in the NRL report which is useful for perspective is that if you take a typical marine mobile communications environment, there could realistically be expected to be several tens of outgoing communications channels (bearing in mind that long distance communications are considered most secure by employing a series of short range ground wave links, each of which is hard to jam.) Calculations were performed based on 64 equal power transmissions being multiplexed through a 1kW PEP system of the ICS3 type. The outcome was that if fundamentals and all up to the eleventh order spurious products are included, (which is where the calculations stopped) 65 percent of the HF band would be occupied by signals at a level at least 40dB above the accepted world average noise level, known to NRL as Quasi Minimum Noise (QMN).

3.5 Wideband Solid-State Non-Lineairties

Wideband, minimum distortion amplifier technology based on solid state devices has been paid great attention by Bell Labs in conjunction with sub-marine telephone cable repeaters. The very high capital cost of laying and maintaining such cables has resulted in much motivation towards research into reliable, optimised systems. The eradication of thermionic valves and the thermal effects that they necessarily involve improves reliability instantly. In 1969, Narayanan analysed several different transistor configurations, looking at single devices and cascaded devices. The object was to find the optimum configuration for long distance telephone repeater amplifiers.
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which have the requirements of good intermodulation performance and reasonable gain. It was soon realised that a single stage would be incapable of achieving satisfactory results and the bulk of the detailed descriptions were of various types of cascade configuration. The relevant conclusions drawn are summarised as follows:-

1) The distortion due to a cascaded stage can be different from either of the two individual stages (either better or worse).

2) Especially noticeable in the case of systems with very many amplifier stages e.g. long distance telephone repeater links, is that third order distortion tends to add in phase as opposed to the power addition of 2nd order products. (This had been observed within Bell Labs almost 50 years previously. No solution is offered.)

3) The common emitter-transformer-common base cascade is the most promising. The transformer is shown to be critical to the modulation performance by optimising the ce-cb impedance match such that the ce stage sees a higher impedance. As it happens using a transformer here raises the gain as well although there may be additional problems in trying to keep the frequency response flat and maintain stability when feedback is applied.

Whether it is as a result of this paper or not is uncertain, but successful use has been made of this configuration in Cable TV (CATV) amplifiers, where very good intermodulation and crossmodulation performances are essential. What is certain is that the SOA solid-state amplifier used for the ASWE report used this configuration because it was recognised as performing well in CATV applications. This was stated in the original description of the amplifier circuit by Boekhoudt of Philips.
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(Eindhoven) in 1976. (fig.3-2) The amplifier was built by MEL and based on a modular system, each module supplying 50 W in Class-A push-pull, the modules being coupled together using hybrid combiners. The reasons given for the remarkable performance of this configuration are:

1) The major distortion causing phenomenon of variation of collector capacitance with collector voltage in the common emitter stage are reduced by a factor of 10 to 1 by loading it with the very low input impedance of the common base output stage.

2) The collector capacitance of the output stage has its effect reduced due to the common base configuration - it results in very little feedback to the input, whereas in common emitter it would be the Miller capacitance, unlike the conventional common-emitter configuration.

\[ RS \]

\[ RL \]

**FIGURE 3-2. BOEKHOUDT'S ORIGINAL CIRCUIT**
There are three established methods for wideband distortion reduction;— predistortion, feedback and feedforward correction.

3.6-1 Predistortion.

Predistortion techniques have been used where amplifier transfer functions have been altered by adding networks before the amplifier such that the combined transfer function of the two networks is more linear than either of them individually. (Some of the effects seen with different types of transistor cascade may be classed as this for instance), although in general the use of synthesised networks has concentrated on reducing third order IMPs (in many applications the principal offending products). This, however, does not necessarily mean improvement of higher order products.

One paper has been seen where, by running a bipolar device at a particular bias point (class-AB with quiescent current at 10% of peak), a cancellation effect has been observed. It was described as the combination of two types of distortion which occur naturally, one near the crossover point and the other near the saturation point and these happen to be phased correctly for cancellation. The device had been measured for changes of phase and gain with power level, and subsequently these characteristics were mathematically analysed to show fair correlation with the observations. It was pointed out that bias network source impedance is important and a special bias circuit was described which has a low impedance up to the maximum difference frequency to be handled by the amplifier. Unfortunately the analysis only extends as far as 3rd order
products and no information is given even about measurements observed of other products.

3.6-2 Negative Feedback

This is the most well known technique of distortion reduction. Its use in electrical amplifiers originated with Black in 1926 and a mathematical analysis of the stability requirements was published by Nyquist in 1932. Many further works have appeared since then, some of which have become classics. The true inventor of negative feedback is more elusive as many mechanical systems and chemical engineering processes can be analysed in a similar fashion. (Steam engine governors, for example).

The chief limitation with negative feedback in wideband systems is that of unwanted time delays in the circuitry which cause phase shifts, whether these be due to physical effects (e.g. transmission delays along conductors or transit delays in active devices) or electrically induced delays (e.g. feedback signals derived from load voltages which themselves are a function of device output current and the (complex) load impedance). In the limit, the well known result occurs that a negative feedback system with sufficient phase shift becomes a positive feedback system, which will in turn become unstable and will oscillate. In real systems the cause of a spurious oscillation may not only be the signal and feedback circuitry, but may also involve unknown signal paths, commonly through bias supply or power supply decoupling components.

While the design rules for engineering feedback amplifiers at high frequencies have long been been defined based upon linear circuit theory, and in practice involve 1) using active devices
Wideband Systems

which do not introduce significant phase shifts (i.e. which may normally be usable to much higher maximum frequencies in narrowband applications), ii) employing compact physical construction to minimise parasitic inductances (which, for discreet solid-state amplifiers, are generally more significant than stray capacitances) and iii) ensuring that load impedances cannot introduce excessive phase shifts in the feedback loop.

Traditionally, the injected feedback signal (whether voltage or current) is some fraction of the output signal (likewise), the fraction being derived by resistive techniques. Recently a trend has emerged to perform "lossless feedback", usually by transformer or directional coupler techniques, figure 3-3. The advantages are:–

i) Improved noise figure owing to the avoidance of resistive (noisy) losses in the input circuitry.

ii) Direct transmission of input power to the output through the feedback network, which adds to the output signal.

FIGURE 3-3. A LOSSLESS FEEDBACK ARRANGEMENT
Wideband Systems

iii) Avoidance of amplifier output power being dissipated in the feedback network. By conservation of energy, it must therefore appear at the output, resulting in increased peak power capability.

Thus three mechanisms for extending the dynamic range are introduced simultaneously. With use of transmission-line broadband transformers,\(^57,58,59\) many octaves can be covered in one amplifier, the limitations in practice being transformer and transistor transmission delays at high frequencies (resulting in instability) and loss of transformer action at low frequencies.

3.6-3 Feedforward Correction

Feedforward correction has been mentioned in one of the earlier references\(^40\) and a recent re-popularising of past work has revealed what appears to be a most powerful and somewhat overlooked tool.

H.S. Black conceived feedforward in 1923\(^17\), four years before his feedback principle and it was patented in 1929\(^60\). It proved impractical though, as gain variations in the valve amplifiers at that time required daily adjustment of filament currents in order to achieve consistent system performance.

Feedforward has since then been re-invented by Pedersen in 1938, whose patent application was rejected however because of a patent held by Macalpine from 1936\(^61\). Pedersen published in 1940\(^62\) on the basis that Macalpine's patent gave no experimental verification or theoretical analysis. Subsequently more work has been carried out by Van Zelst in 1947\(^63\), McMillan in 1951\(^64\), and Schmidt-Bräken in 1960\(^65\) who took out a further patent based on the recognition of amplifier time delay being a correctable
Wideband Systems

function by the use of feedforward.

A major work was produced by Seidel et al in 1968 in which high power (10 W) solid-state pulse amplifiers had feedforward applied to them in order to meet a tight gain and time delay (dispersion) specification over a 40 percent bandwidth centered on 75 MHz and subsequently again at 30 MHz. Intermodulation was not a design criteria in this work and hence no intermodulation performance figures were given. However Seidel went on to apply feedforward to a telephone repeater amplifier over the band 0.5 to 20 MHz. This time intermodulation distortion was all important and it was shown that the original amplifier performance of -65 dB 3rd order IMPs could be improved by 42 dB to -107 dB. Seidel's next project was a microwave Travelling Wave Tube (TWT) repeater amplifier, the performance of which was raised from -78 dB to -125 dB 3rd order IMPs using two feedforward loops.

From this period feedforward was clearly in vogue with appearances in major editorials and at major conferences and also in various descriptive papers. All of these were solutions to multi-signal problems, three for CATV and one for microwave frequencies.

Most recently the NRL have published results of an HF multiplex amplifier with two feedforward loops which have adaptive correction. The amplifiers within the loops are class-A solid-state units coupled in push-pull and subsequently multiple units coupled in phase quadrature. (In the same manner that push-pull results in the cancellation of even order products, so quadrature coupling cancels many odd-order products. The engineering of wideband 90-degree hybrid transformers is not straightforward, however.) The results obtained by the Power Bank
Wideband Systems

are -85dBc for all products with two 100W tones produced from a 500W-peak rated system, i.e. with the peak envelope power at 80% of maximum.

Clearly there is considerable future in the feedforward principle. However the linearity requirements of other components needed in such systems, such as the various couplers, is necessarily leading to the advancing of these other arts as well.

The principle of feedforward is shown in figure 3-4.

A1 is the main power amplifier. This amplifies the signal and feeds its output of signal plus distortion to the system output port via the delay t2 (necessary to match the delay around the correction loop) and the directional coupler C3.

The auxiliary amplifier, A2 has an input, point X, which consists of an attenuated version of the main amplifier output with the source components phased out, i.e. only the amplifier

FIGURE 3-4 A BASIC FEEDFORWARD SYSTEM
distortion products and any amplifier noise remaining. This signal at point X is thus amplified and combined in antiphase with the main amplifier output at the directional coupler G3, the system output now consisting only of the amplified version of the input. As the amplifier noise is also cancelled out, the system noise figure can be lower than that of the original amplifier. It should be noted that all forms of amplifier error are corrected i.e. intermodulation and harmonic distortion, noise, gain and time dispersion.

This idealised system obviously has limitations:

1) If the auxiliary amplifier creates any distortion, it will be coupled to the system output without cancellation.

2) If the first cancellation loop (which reveals the error signal) is out of balance then the auxiliary amplifier may have to handle more power than otherwise, increasing the risk of distortion.

3) The degree of error correction is highly dependent on the accuracy of the second loop. The actual dependence on accuracy in the second loop can be seen in figure 3-5. The resultant signal after feedforward has been applied, R, is taken to be 20dB down on the original distortion component, D. For this the accuracy of the correction signal, C, has at worst case to be within 0.3dB in amplitude and 4° in phase.

4) Reflections back into the amplifier from the load may cause problems also. Any distortion caused by these in the main amplifier output will be cancelled out by the error correction but any distortion actually caused in the output of the correction amplifier will not be cancelled at all.
Wideband Systems

Residual error, $R$, maximum when contributed to by both amplitude and phase errors in $C$.

\[ \text{LOD} = 45^\circ \text{ for worst case.} \]

For $R = 0.1D$ :-

Amplitude error of $C$ \[ \frac{D - 0.7R}{D} = 0.6 \text{ dB} \]

Phase error of $C$ \[ \tan^{-1} 0.7R = 4^\circ \]

**FIGURE 3-5. CANCELLATION LOOP ACCURACY REQUIREMENT**

FOR 20dB DISTORTION CANCELLATION

Unfortunately this problem affects the most elegant property of a feedforward system:- consider a good amplifier working into a perfect load (infinite return loss). Say the amplifier has no spurious products greater than -40dB below PEP. Therefore the correction power required at the output is 40dB below the PEP of the main amplifier. Assuming a 10dB coupling loss in the directional coupler, the required PEP rating of the correction amplifier is thus 30dB below that of the main amplifier, which implies that the apparently formidable gain and phase accuracy combined with very good linearity in the correction amplifier loop are easier to obtain.

Of course in any real system the return loss may easily be only 10dB (e.g. HF marine mobile, VHF land mobile or VHF fixed multi-user broadband antenna systems) especially when long term
corrosion effects are considered. The problem of reflections of main amplifier power in real systems may be solved by the use of an isolator (active or passive) at the output of the correction amplifier.

Applications to TV transmitters may not have such severe reflection problems due to the generally higher levels of antenna engineering found in this field, this being mainly due to the susceptibility of A.M. television signals to ghosting.
4-FOCUSBING ON A TOPIC: ENVELOPE FEEDBACK

4.1 Background

In the original brief for this study, envelope feedback was isolated as a particular aspect of linearity which might be investigated in greater detail. After an initial phase of general experimentation into solid state non-linearities, renewed attention was drawn to envelope feedback and its possible limitations by discovery of a publication which contained a single graph of amplifier phase shift with power level at microwave frequencies.7

The history of envelope feedback was outlined in chapter 1. It appeals because of its apparent engineering simplicity as a means of improving linearity in SSB transmitter output stages. At the time of inception of this project, the Polar Loop technique had been invented (although Cartesian Loop was only an idea in various laboratories at that time). The chief disadvantage with Polar Loop (apart from any commercial considerations) is its hidden added complexity in the need for high quality engineering, especially with regard to thermal matching of balanced components and time delay compensation in the phase feedback path, plus the fact that practical results reveal negligible improvement in the second and third adjacent channels.7

In essence envelope feedback takes the instantaneous peak-to-peak input and output voltages of an amplifier and ensures that their ratio (the amplifier gain) is kept constant. Because the rate of change of RF envelope is related only to the
modulation (related, in fact, to its modulus) and not to the carrier frequency itself, feedback may effectively be applied to the RF circuitry with a feedback loop bandwidth far less than the carrier frequency, which would not otherwise be the case. The envelope feedback principle assumes that gain compression is the only significant problem.

4.1.1 Derivation of IMP frequencies

Gain compression may be shown to produce odd-order IMDs by a variety of methods. As the arithmetic derivation of third order products, for example, is \(2f_1 - f_2\) and \(2f_2 - f_1\), it is natural to assume that harmonics are directly involved in the production of these products in a mixing type action. However, owing to the very complex nature of operation of solid state tuned power amplifiers (discussed later, see chapter 8), attempting to explain the IMDs in this way is tortuous and misleading. Indeed, if frequency mixing is considered, other routes to the production of the any individual IMP must also be considered, involving the separate concept of "frequency mixes" (see appendix 2). With this concept, each distortion signal is arrived at by a combination of mixing actions, which in turn may each involve the fundamental, its harmonics and other signals and their harmonics, an individual fundamental or harmonic being involved more than once (if necessary) in the route to one of the components of an individual IMP.

4.1.2 The special case of the symmetrical spectrum

In the special case of waveforms with symmetrical spectra, a better way of deriving the "close-in" distortion products in a non-linear amplifier is to consider the waveform as suppressed-carrier AM, or DSB. It is assumed that all harmonic components
deriving from the non-linearities have been removed by conventional filtering. In the case of a two tone waveform with two equal amplitude tones (i.e. the standard IMD test waveform, figure 4-1a), the tones can be considered to be the sidebands due to a modulating tone, \( f_m \) of \( 0.5(f_2 - f_1) \) i.e. half the difference frequency (fig 4-1b). Now, imagine modulating the carrier (fig 4-1c) with an infinitely clipped sinewave (i.e. a square wave) instead of this single tone. The well known odd-harmonic spectrum will modulate the suppressed carrier and yield a spectrum just as
Envelope Feedback

If the two tone waveform had been infinitely clipped in amplitude, which (in the narrow band) results in the similarly well known spectrum of bi-phase PSK\textsuperscript{78}. Accepting this relationship, we can see that symmetrical amplitude distortion of an RF envelope (i.e. gain compression) results in the same distortion products as those of symmetrical sinewave distortion of $f_m$, except that harmonics of $f_m$ become the odd-order IMD products of $f_1$ and $f_2$. Although this result is not a revelation, the important point is that product amplitudes involving product relationships such as $(2f_1 - f_2)$ may be calculated without considering $2f_1$ etc.

4.1.3 Even harmonics of the difference frequency.

Initial consideration of this proposed analogy might yield questions pertaining to the presence or effect of even harmonics of the difference frequency, $f_m$. Since in practice these are totally absent, justification for this is as follows:- Were the modulating frequency $f_m$ to be complex and to contain a component of $2f_m$, distortion of the modulating waveform could take one or both of two forms, each depending on the phase of the second harmonic component. One form is an inverse symmetry in time between alternate half cycles, (figure 4-2.a) while the other is an amplitude asymmetry between alternate half cycles (figure 4-2.b).

The mechanism of alternating gain between each envelope cycle may result from the presence of an hysteresis effect or similar hard non-linearity operating at the envelope rate, most probably arising in the collector or bias supplies for amplifiers of the type considered here. This could result in the frequency division phenomenon producing the elements of distortion required for the production of even harmonics (plus a DC component which
Envelope Feedback

2nd HARMONIC
IN PHASE

2nd HARMONIC
SHIFTED 90

FIGURE 4-2. WAVEFORM DISTORTIONS DUE TO 2nd-HARMONIC
OF DIFFERENCE TERM

gives an RF component at the "suppressed" $f_c$). Although such
effects have been observed by the author, they have only occurred
in amplifiers with circuitry containing unintentional
instabilities and hence are regarded as resulting from fault
conditions. As these conditions have always responded to
engineering correction, they may be discounted for the purposes
of this argument.

For a two-tone waveform to be distorted such that successive
envelope zeros are moved alternately back and forth in time, a
distortion mechanism is required capable of storing and releasing
energy repetitively at the correct part of successive envelopes.
No such mechanism for this exists in normal amplifiers. Thus
there is no reason to expect to find components relating to even
harmonics of the difference frequency in this analogy.
4.1.4 A digression

An interesting digression at this point is to consider the standard practice in the USA to deliberately apply asymmetrical modulation distortion (therefore involving even harmonics of the modulation signal) in AM broadcast transmitters at MF. This principally allows up to 150% modulation depth (limited by FCC regulations) without incurring hard non-linear distortion in receiver detectors (which are normally asynchronous peak or square-law devices), and hence a greater audio output for a given RF input level can be achieved. This has the subjective effect of making a given station seem stronger than another (even though it may be weaker in field strength terms) thus improving the "rating" performance, the all important statistic in the ultimately commercial US broadcasting industry.

The effect on the transmitted spectrum is to place more energy further out from the carrier than without such distortion. This also has another subjective effect:-- not only is there more sideband energy in a given transmission, but because this extra energy is mainly located further away from the carrier, upon tuning toward such a station it sounds considerably louder as a result of these two mechanisms.

4.2 Methods of applying envelope feedback

4.2.1 Gain correction

Gain compression may be compensated for in a number of ways:

i) Adjustment of some parameter of the amplifier itself, e.g. adjustment of bias conditions (so called "bias pumping" or
Envelope Feedback

"sliding bias") or by power supply modulation.

ii) Gain adjustment in a previous stage, e.g. in the driver stage or by a variable attenuator between the driver and the power amplifier.

iii) Overall system gain adjustment by feedback to some early stage, e.g. an IF stage.

iv) Adjustment of a post amplifier stage, e.g. a variable output attenuator.

The first two methods are the most common found in the literature\textsuperscript{79,80,81}. In depth studies of these techniques and their relative merits and limitations are rare (which is a significant reason for the funding of this study).

Treatment of one isolated stage of amplification is fine in its own right, however all real systems have multiple stages and the interaction between multiple power amplifier stages is also not well documented. It is widely known that a) multiple stages can produce different output spectra than either stage individually and b) the voltages and currents at the interstage may have IMD spectra which differ substantially from the final output spectra. These two statements are true for both wideband\textsuperscript{82} and narrowband situations\textsuperscript{83}. The implications of this to envelope feedback are:-

i) Multiple stage compensation methods may have greater complexity than those for single stages.

ii) The interstage measurements imply interaction between stages. This may also be viewed as dynamic loading for
inputs and outputs of all stages, or alternatively as a natural form of pre- (or post-) distortion.

4.2.2 Methods of envelope detection

The two principal methods of detection are the use of diodes (either singly or in bridges) as peak detectors and synchronous detectors (using a variety of different switch elements). The most stark difference between these is the low signal performance: peak detectors do not work at all below the diode 'cut-in' voltage. Peak detectors may be modified to some extent by being biased to the edge of conduction, and in the limit may be used on the non-linear conduction edge, the (exponential) non-linearity providing the rectifying function. This mode represents a third detection method often termed 'diode bend' or 'square law', the latter owing to that component in the series expansion of the transfer characteristic which is primarily responsible for the production of the baseband frequencies.

Insufficient harmonic filtering at the amplifier (input or output), and the subsequent waveform distortion, will affect the performance of peak detectors. If a synchronous detector is used, even-harmonic RF signal distortion will be suppressed to a great extent, but odd-harmonic effects will not be so significantly reduced. The resulting effect will be the production of an error signal which will be corrected (falsely) by gain adjustment. Hence the wanted RF signal will become modulated at the envelope rate producing sidebands which will be interpreted as IMPs.

The insensitivity of a synchronous detector to the various harmonics can be seen in figure 4-3. Fundamental power is always of one polarity, while the second- and all higher even harmonics produce waveforms with an average of zero, and third and higher
Envelope Feedback

odd harmonics produce waveforms which do not completely cancel. In practice, distortions in the rectifier driving waveform and device symmetry limitations typically result in measured even-harmonic suppressions of 20dB.

![Waveform Diagram]

**FIGURE 4-3. SYNCHRONOUS DETECTOR WAVEFORMS**

4.3 AM-PM conversion

If, as the drive power to an amplifier is altered, the relative phase between the input and output signals vary, AM to PM conversion is said to take place.

4.3.1 The AM-PM problem

This effect implies that any amplitude-varying signal will have introduced to it some phase modulation after passing through an amplifier which exhibits AM-PM conversion. PM signals also
Envelope Feedback

exhibit sidebands and these appear as IMPs. Hence envelope feedback alone cannot remove IMD completely in this case.

The relationships between the angular modulation of a vector and both i) the in-phase and ii) quadrature modulation components subsequently imposed on that vector are non-linear (figure 4-4). From this, and despite assuming a linear AM-PM conversion function (which is not generally true), the IMFs due to AM-PM conversion tend to have more power in the higher order products than for gain compression (AM-AM conversion) alone. For SSB transmitters this is, in practice, very significant. (It is anomalous that non-linearity specifications are usually for low order products alone).

FIGURE 4-4. PHASE MODULATION ANALYSIS
Envelope Feedback

In the broadband TWT transponder amplifiers used in microwave satellite communications, AM-PM conversion is a much more recognised problem, and typical AM-PM conversion characteristics are known and quantified with the parameter $k_p$, measured in degrees per decibel. In practice this is expressed as a plot against input power.

4.3.2 Quantifying the problem

Only one report containing quantitative information on AM-PM conversion in solid-state amplifiers has been found, and this was in a microwave application. As it is clearly going to be of great importance to envelope feedback, typical AM-PM characteristics for HF and VHF amplifiers must be established (because it is at these frequencies that interest in single channel SSB transmitters exists).

An initial experiment was set up as shown in figure 4-5. The amplifier circuit figure 4-5(b) (i.e., impedance matching and bias components) was extracted from the application note for the device (Mullard BLW31) in order to create typical operating conditions.

The amplifier was adjusted sequentially at the output for maximum output power, and at the input for maximum return loss (tending to 1:1 SWR). The phenomenon of the unity match condition being unattainable was experienced. This is due to the presence of RF harmonics generated in the test amplifier appearing at its input. The solution is to employ a tuned indicating device as the reverse power detector, instead of the more usual diode detector method found in commercially available SWR indicators.
FIGURE 4-5 INITIAL AM-PM MEASUREMENT
The extent of the phase deviation is shown on figure 4-6 and agrees well with the plots originally published by Sechi for a microwave amplifier\cite{119} and clearly emphasises the need for more detailed investigation.

4.4 Experimental objective:-- The control of Phase Modulation

As the extent to which envelope feedback can combat distortion is going to be limited at least by the AM-PM conversion characteristics of the amplifier, there is a need to establish a method for controlling the phase of the amplifier. Hence experiments were carried out to find the effects of the variation of base-bias and supply voltage on the phase shift of the amplifier.

Considering i) the historical significance and ii) the comparative engineering simplicity of "bias pumping" techniques, these were investigated first. Little useful correlation was
Envelope Feedback

found however. The same cannot be said for supply variation.

By simultaneous manual adjustment of supply voltage and RF drive power, a region of operation was discovered where phase could be kept constant for any value of drive power by the selection of the correct supply voltage. Moreover, when a plot of gain was made over 10-1 power range, this too was found not to vary to any great extent.

From these results three fundamentally important implications can immediately be drawn.

1) Controlling the supply voltage from the integral of the phasemeter output will control AM-PM conversion and gain compression simultaneously over a wide range of input power.

2) Controlling the supply voltage purely from envelope feedback may control AM-PM conversion also, but only over a limited region of operation.

3) Controlling the amplifier power supply to manipulate gain and phase in the way that is required (i.e. reduced voltage at lower powers) will also improve amplifier efficiency. Improved total transmitter efficiency may be obtained if a high efficiency variable power supply is used (e.g. switched mode).

4.5 Construction of an envelope feedback amplifier.

To assist with testing these hypotheses, the amplifier of figure 4-5b was modified to include a collector supply series modulator and a pair of envelope detectors. The circuit diagram is shown in figure 4-7. The RF section is the same as used for the original measurements.
4.5.1 Bias supply details.

As is normal practice, the bias supply is required to provide a constant voltage which tracks with the temperature of the RF power transistor. The usual technique is employed here, which is to thermally couple a temperature sensor (a diode with a continuous forward bias current) to the case of the transistor. This has its limitations: the RF transistor junction temperature is the relevant parameter, and due to the construction of the device, a case mounted sensor will not reach the junction temperature. Using the concept of thermal resistance, the thermal properties of a device mounted on a heatsink can be modelled as an electrical circuit, see figure 4-8. The failure of external sensing to measure the true temperature can be seen from this.
However, as the thermal resistances are nominally linear, this can be compensated for and hence the use of the two diodes (D3, D4) in series. The modification from standard two transistor method of figure 4-5b to the use operational amplifier circuitry is for ease of adding control signals to the bias voltage.

It is normal for device application information to include the thermal resistance data. However, it is less normal to include thermal capacities as well (which can be modelled as electrical capacitors from thermal resistor nodes to an imaginary infinite sink, or ground), and which could yield information regarding thermal risetimes. Thus true dynamic compensation is difficult to design for. The usual methods for control loop compensation may be applied, using empirical measurement techniques (such as square wave injection), with one exception. This is that the bias supply is a voltage source and the bias
current drawn has an exponential relationship to this voltage. Thus thermal time constant compensation (which is likely to be carried out linearly in the voltage domain) must be passed through a logarithmic conversion. Difficulties enter here due to other thermal effects influencing this conversion. Hence dynamic compensation is not often found in practice.

The bias current drawn is also related to the magnitude of the input signal power, owing to the rectifying nature of the base input circuitry. Thus bias current has frequency components in excess of the channel bandwidth, and therefore the bias supply response time (which can alternatively be viewed as AC output impedance) must be short (low at high frequencies). Failure to achieve this results in extra distortions due to dynamic changes in bias conditions.

4.5.2 Envelope detectors and feedback circuitry.

Because i) the project brief was to investigate envelope feedback and ii) to keep the amplifier as simple as possible, diode detectors were used to detect the envelope of the signals at input and output. Other options include the use of synchronous detectors or alternatively, the construction of a phase feedback system using phase detectors. It may be considered that the use of diode detectors, with their inherent non-linearity at low power levels, may be adding complexity. With this in mind, the opposed detector plus attenuator technique was employed, as was also used by Terman and Buss\textsuperscript{23} (see chapter 2), the detectors being used at the same power level and such that their non-linearities cancelled.

The final error signal results from current summing at the virtual earth node of the inverting amplifier IC1, figure 4-7. At
Envelope Feedback

progressively lower power levels, where no error signal will be produced and the dynamic (small signal) diode impedance becomes very large, the amplifier conditions become increasingly controlled by pre-set components in the power supply modulator i.e. VR1.

(The term modulator will be used synonymously with variable collector supply from this point onwards. It is not meant at any point in the thesis that any component of the baseband signal is introduced directly into the modulator. The justification for the use of this term is that although the supply modulator is not being used to add some unrelated modulation to the amplifier input signal in the traditional sense, it is being used to control gain, and adjustment of gain conventionally adds modulation.)

Conventional negative feedback amplifiers control gain by feeding back a known proportion of the output signal to the input in anti-phase with the current input signal. If the gain of the amplifier without feedback is sufficiently high, the gain of the final amplifier is controlled by the feedback proportion. For an envelope feedback amplifier this is not the case. The amplifier gain is measured by the difference of the (suitably scaled) input and output signals. The magnitude of this difference is dependent on gain error and on the magnitude of the two signals. For very small signals the error signal tends to zero, even for large gain errors. This effect is increased further by the use of diode-peak-detectors. Final description of envelope feedback mathematically or as a control system is made more complex still by effects to be mentioned, and to be discussed later.
4.6 Initial results

Setting up the feedback amplifier may be tackled from two directions:– Either by minimising the phase shift over the working signal range or by optimising the spectrum (gain must be optimised by the very nature of the system). It was found that first optimising the phase and then the spectrum was best. Interaction of spectral behaviour between bias, quiescent supply voltage and amplifier gain (selected by adjustment of the attenuator before the output detector) was considerable. However, after adjustment the spectral performance shown in plate I was obtained. This was quite stable in its own right, being repeatable on a day-to-day basis. Phase performance under these conditions was as shown in figure 4-9, and can be seen in comparison with figure 4-6 to be considerably linearised up to the saturation point.

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**FIGURE 4-9. PHASE SHIFT OF ENVELOPE FEEDBACK AMPLIFIER**
Envelope Feedback Amplifier Spectrum with Associated Operating Waveform

Typical spectrum of an optimised envelope feedback amplifier. The most significant improvements lie in the medium order components (i.e. 5th, 7th, 9th, etc). The oscillograph shows the collector supply voltage (centre line is zero) with the suitably scaled RF output waveform superimposed. Different optima result in collector supply waveforms which differ at the low power point of the envelope, typically remaining at approximately 2V.
Envelope Feedback

The important points arising from these results may be summarised:

1) Simultaneous phase and amplitude distortion cancellation has been shown to occur in single ended, tuned, bipolar power amplifiers at VHF and HF in different devices.

2) From the close correlation between measurements of phase shift and those published of a microwave bipolar amplifier, it can be inferred that this cancellation effect may be found at microwaves also.

3) Despite the use of simple diode peak detectors, 3rd order IMD performances of the order of 40dB can be achieved. Medium order IMFs are also particularly improved.

4) The upper power limit of the amplifier is no longer set by device saturation in the normal sense, but by maximum available power supply voltage. This observation was not tested to the point of destroying the amplifying device.

5) The maximum (low power) gain is somewhat reduced over that of the non-feedback amplifier (3-5dB), i.e. the amplifier is being operated in some degree of gain saturation.

4.7 Operational difficulties:-- the need for more data.

The interactions which exist between the various parameters of the total system are sufficiently complex as to defy detailed description or explanation with currently available information, but can be briefly summarised:
Envelope Feedback

1) If manual adjustments of variables are performed according to some subjectively assessed iterative algorithm, different optima can be found for different algorithms. e.g. low order products may be optimised at the cost of high order products and vice versa. Overall optimisations do seem to be available, but only after extended periods of adjustment.

2) Monitoring of input SWR during adjustments showed curious variations.

3) As the quiescent bias condition found necessary for optimum performance is relatively low (<1% of maximum signal current), the large signal bias current requirements cause difficulty for the bias supply. This, of course, is principally an engineering difficulty.

The experimental results outlined in this chapter represent a point of reference for the project. The remainder of this thesis is devoted to detailed measurement, analysis and assessment of the validity and extent of the results and claimed benefits.
5. DETAILED MEASUREMENTS:— PHILOSOPHY AND HARDWARE

5.1 Required parameters

From the experimental results of the last chapter, and in particular from the interactive nature of the various adjustments on the envelope feedback amplifier, consideration was given to the accurate measurement and multi-dimensional presentation of the amplifier characteristics. The following parameters were specified as being essential for adequate analysis of the amplifier conditions:

Controls;

1) Collector supply rail voltage control.
2) Base bias voltage control.
3) RF signal source voltage.

Measurements;

1) Supply current.
2) Bias current.
3) Output power.
4) Input/output relative phase.
5) Input reflection coefficient.
6) Input reflection phase angle.

In addition, for checking purposes during automated measurements.
5 - Measurements preparation

7) Supply voltage at amplifier.
8) Bias voltage at amplifier.
9) Input power at amplifier.

5.2 Practical limitations: - available test gear

Although some test gear can be constructed, the presence or absence of major items is a controlling factor. Equipment available (some only later in the project) on a more or less permanent basis included:

2) Racal IEEE-488 True RMS level meter 9303.
3) Marconi spectrum analyser TF2370.
4) HP vector voltmeter 8405A.
5) Commodore computer (PET) 3032 series with dual disc drive.
6) High power load.

Equipments required to transform this into a workable system were:

1) Programmable DC power supplies
2) Various signal routing switches and terminations (50-ohm co-axial system)
3) Various DC switch conditions
4) 12-bit IEEE-488 multi-channel A-D and D-A converters, to interface the above items with the computer.
5) Various RF signal couplers and power splitters.

The Marconi spectrum analyser has an upper frequency limit of 110 MHz. As it is very useful to have to hand an indication of amplifier harmonic performance as well as close in IMD levels,
FIGURE 5-1 ATE SYSTEM BLOCK DIAGRAM
the frequency of 29MHz was chosen for the experiments.

The final system block diagram is shown in figure 5-1.

5.3 Measurement system details: Analogue

5.3.1 RF Components

The 2019 signal generator is fitted with the general purpose interface bus (IEEE-488, see appendix 3) and can thus be programmed directly from the computer. The generator has a maximum power of 13dBm (20mW). Hence a post amplifier was used (This was a highly linear breadboard model generously donated by J.Ling of the Mullard Application Laboratories, Mitcham). This brings the maximum available power in excess of 1W. The attenuator between the generator and amplifier serves two roles; that of

i) Allowing a round number attenuation between generator reading and actual output. (A total error figure of 20dB was chosen, being easy to remember and resulting in a x10 error in voltage settings).

ii) Improving system noise performance. Due to the use of wideband amplifier techniques in the signal generator, the signal to noise (floor) ratio at the generator output is poorer at lower signal levels. This otherwise leads to reading errors owing to the wideband nature of the Racal level meter.

The attenuator after the post amplifier is to ensure;

i) Improved return loss between this amplifier and the Amplifier Under Test (AUT). This minimises a) the risk of incorrect readings of forward power into the AUT and b) the
reflection at the post-amplifier output of IMPs transmitted backwards from the AUT as a result of dynamic variations of AUT input impedance. (a) is the same effect as (b) except as observed under static, signal-up conditions.

ii) A worst case 6 dB return loss output load for the post amplifier, which might otherwise have destructive results.

The directional coupler was constructed in a co-axial manner using readily available materials. The design works by a combination of inductive and capacitive coupling between the coupled line and the field in which it is mounted. The design therefore has a frequency sensitive coupling coefficient. However, experiments with directivity revealed no anomalies which would detract from the coupler's usefulness. The principal reason for using this design is the total lack of non-linear components, i.e. ferrites, thus avoiding the subsequent possibility of introducing extra distortions.

Coupled forward power is used as a phase reference for all phase measurements and was also initially used as a measure of input power to the AUT. Due to interference effects from the vector voltmeter creating an artificially high noise floor (covered more fully in section 5.5.2), this arrangement was later modified as shown with the level meter being fed from a separate, amplified sample of the generator output. This amplifier itself has a wide bandwidth, hence the use of a low pass filter to improve noise performance. Output from the high power attenuator was split into two with a hybrid power splitter. (The term hybrid is used for many engineering applications, in this instance it means the use of coupled voltage and current transformers.) As shown, the splitter feeds the second input of the level meter and the second input to the vector voltmeter. Justification for the use of a device containing ferrite components here is that i) the
load impedances are known and constant and ii) the power at this point cannot exceed 40mW, measurements proving the hybrid to be free of significant distortion components at this power level.

As the vector voltmeter may also have its second input switched to the reflected power port of the AUT input directional coupler, both the hybrid splitter and the directional coupler must have the facility for being terminated in 50-ohms when not being used to feed the vector voltmeter, otherwise the directional coupler and hybrid will not work as desired because the unterminated output will reflect the incident power back through the device. The switching and termination are achieved by two relays within the co-axial switching unit.

RF power was measured using the Racal 9303 level meter. Particularly useful facilities were:

1) Two measurement heads which could be switched over the IEEE-488 bus.
2) Programmable averaging time.
3) Ability to calibrate out unknown attenuation factors of up to 40dB, i.e. the high power output attenuator and the input amplifier feed arrangement.

The Marconi spectrum analyser has a pen-plotter output and this, combined with its 90dB dynamic range (for spot source measurements in narrow bandwidths), makes a useful filtered voltmeter. For measuring the fundamental component of the reflected input power, it proved essential.

5.3.2 Power sources

The voltage sources for supply and bias were series
regulators with inputs programmed by outputs of the D-A converters. They included current limit circuitry and are designed for frequency compensated operation into the expected load impedance, which is highly capacitive due to the large decoupling capacitors present at the amplifier power ports. Although the AUT was to be tested at static RF power levels, the supplies were frequency compensated to allow dynamic testing with the least amount of re-connection.

Current outputs were provided in a form suitable for the A-D converters by the use of carefully balanced differential amplifiers.

5.3.3 Equipment interfaces

Equipment requiring custom interfacing to the computer system were as follows:

1) Vector Voltmeter. Outputs of amplitude (1V=FSD) and phase (+0.5V=+FSD) were provided. Simple operational amplifier circuits process these signals to suit the A-D converter inputs (0-10V, 2.5mV unit step). Mechanically switched amplitude ranges limited the usefulness of the amplitude output. However, setting the phase range to +180° allowed the computer to measure all possible angles with a repeatable resolution less than 0.5°.

2) Spectrum analyser. An output was available over the range -100dB=0V, 0dB=2V.

3) Post amplifier switching. During setting of the bias conditions (see section 5.4.3) RF power needed to be switched on and off for a very short period. Bus operations
5 - Measurements preparation

to the signal generator limited the minimum period to approximately 0.5 seconds. This is two orders of magnitude too long and thus a solid-state power supply switch was designed for the post amplifier.

4) Co-axial relay switching (CR1 & 2).

5.4 Measurement system details: Digital.

Initially the system employed some already available 8-bit IEEE-488 converters. Early measurements with these suffered from inaccuracies, principally due to excessive bias voltage and collector current increments. After some effort was expended to alleviate these problems i) with logarithmic voltage converters for the bias supply and ii) measurement windows for collector current, a 12-bit system was designed and constructed. The results proved the effort to be worthwhile.

5.4.1 Data aquisition.

The IEEE-488 system has 8-bits reserved for data in each byte. Transferring 12-bit information thus requires at least two bytes. It is normal to use more than this, computers then being able to print strings of ASCII characters, one per byte, each one representing individual decimal digits or control code alphanumerics. A high data transmission rate is imperative in this case and thus the system was optimised for this, despite the increased software complexity needed to encode and decode the data. The block diagram of the data conversion unit is shown in figure 5-2. The input and output channel capacity was chosen according to the following criteria (in order of importance):
5 - Measurements preparation

1) Known immediate measurement requirements plus a margin for possible extra channels

2) Two-byte addressing limitations with 12-bit data.

3) Popular CMOS device architectural limitations, e.g. analogue switches.

4) Maximising system usefulness within the above constraints.
5 - Measurements preparation

Certain seemingly trivial features were included, principally in the area of input channel and voltage indicators. In practice, these are essential time-saving devices which were not present on the original 8-bit model; during software development the program may halt for many reasons. To know which channel the converter is addressed to plus the approximate voltage on that channel may otherwise take considerable time, or be impossible to establish.

5.4.2 Control system

The only computer available to the project on a permanent basis which had real-time and IEEE-488 controller facilities was a Commodore PET. This has a 32k memory and works in BASIC. The IEEE-488 functions are not fully implemented, which leads to limitations and to problems with some types of equipment. The BASIC is interpretive i.e. the machine code for the central processor is generated as the program is running. Compiled BASIC could have been used (yielding considerably reduced run-time) but the software generally required modification at a rate which would have caused the time lost during compilation to be excessive.

5.4.3 Data storage

The decision to display the amplifier measurements in a three-dimensional format implies the use of a mainframe computing facility. The University PRIME computer network has the necessary graphics program facility and is nominally real time, however the demand for terminals is high during the day, with response times
5 - Measurements preparation

often very poor. For this reason and also the complexity of available programs which employ the Rutherford File Transfer Protocol (FTP) supported on the PRIME, the data from the measurements was stored locally, and transmitted outside the experimental period to the PRIME system for processing.

The file format used was such that for each set of measurements under a given set of conditions, the data were stored sequentially with each value being separated by a comma. The sets of values for new conditions were each separated by line-feed. Thus listing the data by all conventional means (i.e. on a VDU under file editing control or printed out on paper) results in the set of values for one condition being shown on one line. Reading files formatted in this way into a FORTRAN program is also most convenient.

Care was taken to ensure that duplicate data was stored in more than one place, in case of accidental erasure.

5.4.4 Control algorithms

Owing to its cyclic and repetitive nature, Automatic Test Equipment software lends itself to a subroutine based construction with nested conditional loops. The complete flow diagram is shown in figure 5-3, and is mainly self-explanatory apart from the following points.

1) Measurement and setting of bias conditions. This was made complex by heating effects within the transistor. The bias supply is a programmable voltage source, but the current drawn by the semiconductor junction is strongly temperature dependent and follows the well known equation 5-1.
5 - Measurements preparation
The temperature $T$ is the junction temperature (in Kelvin) which is controlled by device dissipation. The time constant of the junction temperature is a non-linear function due to multiple thermal mass elements (CR values in the thermal equivalent circuit), as described in section n-n. Device dissipation is most strongly a function of DC input power i.e. supply voltage multiplied by DC input current, but also involves DC-RF conversion efficiency which varies with the signal level. The problem is that owing to the rectifying nature of the transistor input, bias current cannot be measured in the presence of the RF signal which controls the DC input power. Since the device dissipation is also a function of bias conditions (especially at low power levels), the biasing arrangement is made potentially unstable, commonly termed thermal runaway. Stability was ensured by appropriate software control loops.

Experiments were carried out to establish a measure of the thermal time constant found in practice. Since collector current is related to temperature by almost the same relationship as base current, plate II shows oscillographs of collector current which illustrate decay in the junction temperature at two different time scales and at two different bias current settings. From the graphs there is a time constant of the order of 5 seconds and a second superimposed of around 10 mS. Not shown is a much longer time constant due to the heatsink-to-ambient relationship. This is, in this case, of the order of minutes, but in some situations may not be significant at all e.g. if forced air cooling is applied.
Four oscillographs of collector current to illustrate the effective thermal decay time constants acting on a Power Transistor Junction. In each case the RF waveform (lower trace) shows when the amplifier current was reduced from 4A to the quiescent value. The two time constants of approximately 5mS and 10S can be observed on the collector current (upper trace). The centre line is zero collector current.
5 - Measurements preparation

The junction temperature prevailing during normal amplification is dependent on the nature of the signal being amplified. The 10mS time constant will have the effect of altering (in a retarded fashion) the bias conditions during and immediately after abrupt changes in power level. In solid state UHF TV transmitter amplifiers (where device geometries may create even shorter thermal time constants) the high power synchronising pulses cause IMD variations during the rest of the line period for this reason.

In order to achieve consistent results, the 10mS time constant is allowed to decay almost fully before the current measurement is taken. One instrument on the bus system requires several seconds of steady signal to make its measurement. For this reason the bias setting subroutine allows a considerable period for thermal stabilisation, measuring bias current in a minimally short break of RF drive power.

2) Accurate computer controlled measurement of RF level using the Racal 9303 requires care. The difficulties arise mainly from the IEEE bus interface option. The IEEE bus is not implemented identically or fully by any equipment used on this project (especially including the PET bus controller). Driving the Racal instrument from the PET gives particular problems. Many weeks were taken up learning how to take measurements reliably, involving use of logic analysers and extended checking routines in the PET software. Also, due to timing problems in the 9303, the PET software contains many empirically derived wait loops in order not to address the 9303 when it is engaged with housekeeping functions. Failure to take this precaution resulted in many mis-readings and occasionally in total system lock-up, which could only be recovered by terminating the program. With measurement programs which take tens of hours to run, this proved
5.4.5 Software philosophy

For ease during system development the procedure for each measurement or control function was allocated a number block, standard entry points to a given block yielding standard modes of execution, e.g. entry at the beginning of a control block (xx00) results in a prompt for the required output voltage, while entry at a later point (xx20) inputs data with no prompt and entry at the midpoint (xx50) assumes that the variable is already assigned with a value and simply executes the function. Each subroutine returns successfully regardless of entry point. With some of the more complex routines such as signal-up bias condition setting with thermal delay (5500), this feature proved singularly useful. The screen editing feature on the PET allows real time execution of single line programs (e.g. groups of subroutines all separated by colons) without the use of the RUN command, which would otherwise reset all variables to zero.

5.5 Systematic errors

In scientific work based on measurement, it is imperative that the measurements do not contain large errors. Automated measurement systems are perhaps the most vulnerable to systematic error. Much time has been spent in repeated experiments simply in order to gain perception of and confidence in the reliability and reliable accuracy of the system.
5 - Measurements preparation

5.5.1 Precautions with settling time

As has already been mentioned the Racal level meter requires some time to settle before it gives valid readings. The averaging time was set to 4 seconds. This gives a good compromise between measurement noise and time consumed during measurement.

The HP vector voltmeter requires several hundred milliseconds to stabilize within the resolution of the A-D converter.

5.5.2 EMC

In a project working with large RF signals, problems with unwanted signal injection are rife if precautions are not taken. The principal methods of avoiding RF problems are:-

1) Good decoupling of all power rails. This includes recognising that real capacitors have some residual inductance associated with them and must be chosen carefully that the resulting series resonance properties are used to advantage and not unwittingly as RF chokes. Paralleling of multiple values of capacitor is a useful technique, providing care is taken with the parallel tuned circuits which result.

2) Avoidance of stray inductances i.e. any conductor which is unwittingly carrying RF current.

3) Use of electrostatic screens.
5 - Measurements preparation

4) Correct use and (electrical) termination of co-axial cable and connectors, plus continuous caution in their procurement, assembly and maintenance.

Problems were encountered with interference between the HP vector voltmeter, the Racal level meter and the spectrum analyser. The sensing heads for both the Racal and HP instruments use the sampling diode bridge principle. At low measurement levels the noise emitted by these bridges can swamp the wanted signal, especially if broadband measurement are being made. The system configuration was altered on account of this, and the Racal head for input power sensing was fed from its own sample directly from the signal generator instead of from the directional coupler at the amplifier input. This was only done after measurements proved beyond reasonable doubt that the post amplifier was not likely to traverse into saturation. Spectrum analyser problems were solved by reducing the measurement bandwidth.

The IEEE-488 bus carries multiple, high frequency, TTL-derived square edged waveforms between several of the instruments on the system. Although the bus cables were all screened, earth loop and other interference effects were found to exist within the system. The worst affected, and most offending instrument in this respect was the Racal level meter. This instrument was found to interfere with itself even when measuring its own internally generated calibration signal as soon as the bus connector was inserted. Indeed, simply connecting the co-ax outer of the sense head to the equipment ground was sufficient, with all other bus instruments turned off.

The problem was improved to some extent by modification of the internal earthing arrangements of the meter. The bus interface
5 - Measurements preparation

card has been designed to be retro-fittable and, despite being factory fitted in this instance, poor earthing techniques appear to be the principal cause of the interference.

With the benefits of this hindsight, the 16-channel A-D interface unit had all bus signal and control wire pairs bifilar wound on ferrite suppressor chokes at the design stage.
6. DETAILED MEASUREMENTS: INITIAL RESULTS

6.1 Data presentation techniques

Earlier work has shown that the amplifier characteristics vary interactively between a number of parameters. Communicating such information directly is difficult as many dimensions are required. For this thesis, presentation is mainly in three dimensions (i.e. one variable against two parameters), either as isometric projections or in contour. Such presentation is made readily available by the GINO graphics packages, written at the University of Salford Computing Laboratory under Crown funding. The 3-D algorithms were derived from work generated during the early 1970's at the National Physical Laboratory. The GINO packages are accessed as FORTRAN subroutines.

All the plots for this chapter are performed on the same device (BLW60) and at one frequency (29MHz) and, in most cases, under identical tuning conditions. The validity of this approach remains questionable when considering very fine detail. However, this is justified by the successful implementation and similar (one dimensional) plots made on the original 144 MHz amplifier using a BLW31.

6.2 Initial plots.

Conventional linear amplifier operation is at fixed supply conditions, and thus normal operational characteristics can be extracted from the following multi-dimensional presentations by using only that data from along a fixed bias current or supply voltage line, say, that for 500mA in the case of plots against bias or 12V in the case of collector supply.
Measurements Results

**Figure 6-1A** Collector Supply Current Versus Bias and Drive

**Figure 6-1B** Power Efficiency Versus Bias and Drive
Measurements Results

As in chapter 4, quiescent bias current is the first parameter to be tested. Figures 6-1.a and b show respectively contour plots of supply current consumption and power efficiency variation with bias current and drive, the supply voltage being fixed at 10V. It can be seen that at very low drive powers, efficiency is remarkably constant, despite the change of the supply current drawn. This implies a change of output power and therefore of gain which is proportional to supply current.

6.3 Main results

6.3.1 Gain and Phase versus Bias

Variation of gain and phase with RF drive and bias are shown in 3-d and in contour in figures 6-2.a to d. The gain plots clearly show increasing crossover distortion at lower bias and drive levels. For operation below saturation, the bias point for optimum gain linearity can be seen to be about 400 mA. On the other hand, the phase contour plot shows increasing distortion (AM-PM conversion) at higher bias levels as well as at lower ones, the optimum point being about 70mA. Hence an anomaly exists as to the best bias condition to choose for a conventional amplifier.

Plate III show spectra for the same amplifier under both of these two bias optima with both noise band and two-tone test waveforms. The low bias spectra (phase optimised at 70mA) show reduced low-order (close-in) components over those for high bias (gain optimised at 400mA). But note also that for low bias, the slope of the IMD products with product order is reduced.

[In order to achieve the same display height for the noise band case, the analyser is working at a higher peak signal level]
Variation of IMP slope with bias conditions
Noise Band and Two-Tone Tests

For conventional bipolar linear amplifiers the quiescent bias current affects the rate with which the higher order products decrease in amplitude. It is common practice for IMD specifications to be given in terms of 3rd order products alone, under which circumstance it is clearly beneficial to measure at the lower bias. Unfortunately this detracts from the ability to judge adjacent channel performance from IMD specifications.
Measurements Results

![Graph A: Gain variation versus bias and drive in projection and contour](image)

![Graph B: Gain variation versus bias and drive in projection and contour](image)

**Figure 6-2. Gain variation versus bias and drive in projection and contour**
Measurements Results

**Figure 5-2C & D. Amplifier Phase Shift Versus Drive and Bias**
Measurements Results

and hence the apparently worsened noise floor, which limits the measurements. Spectrum analyser close-in dynamic range is almost universally limited by local oscillator phase noise and the associated reciprocal mixing effects.

Since, for successful commercial implementation, SSB communications equipment is expected to have a dynamic range of 90dB or greater, and since there is no reason to expect that the IMD slope may not be extrapolated beyond the measurement system limitations, it may be inferred that, apart from in the immediately adjacent channels, lower bias current leads ultimately to worse adjacent channel performance. From this, the fact that the IMD specification of SSB transmitters is normally based entirely on low order performance may therefore be seen to be misleading. A method based on the IMD magnitude and on the description of slope with frequency would seem to be more appropriate.

Confirming the speculative results of chapter 4, the use of bias voltage for feedback control is thus observed to be impossible for both phase and gain simultaneously, for if bias is used to control gain, phase would remain uncontrolled and require a subsidiary control loop, or vice versa. Under all circumstances, control into the saturation region is not possible, e.g. in figure 6-2.b, in the range bounded by all drive voltages greater than 3V, the gain is progressively decreasing and is relatively unaffected by the quiescent bias.

6.3.2 Input reflection versus bias

Input reflection is a measure of input impedance and as such is of interest because a dynamic input impedance at one stage represents a dynamic load impedance to a previous stage. This may
Measurements Results

RETURN LOSS

(×10=dB)

x100=BIAS (mA)

RF DRIVE (V)

A

B

FIGURE 6-3A & B. INPUT REFLECTION MAGNITUDE VERSUS BIAS AND DRIVE
Measurements Results

RETURN PHASE
(x100=DEG)

x100=BIAS (mA)  RF DRIVE (V)

RETURN ANGLE VERSUS BIAS AND DRIVE

FIGURE 6-3C & D. INPUT RETURN ANGLE VERSUS BIAS AND DRIVE
Measurements Results

be seen as a distortion mechanism in its own right when a purely resistive source is used, apart from any complexities arising in driver stage performance due to variations in load impedance. It has previously been implied that these complexities can sometimes lead to output IMD performances which are better than the interstage values, owing to distortion cancellation. Thus it is relevant to present the input match data.

Figures 6-3.a to d show the input reflection magnitude and phase versus drive and bias. It would appear significant that there is some correlation between the input/output phase contours (fig 6-2.d) and the input reflection phase contours (fig 6-3.d). The conclusion is that phase and input impedance may, therefore, be simultaneously controlled over an extended region by bias pumping.

6.3.3 Gain and phase versus supply

As discussed in chapter 4, the feedback amplifier controls collector supply voltage in order to keep gain constant. Observation of the feedback amplifier operational waveforms (plate I, chapter 4) show that the control voltage is, except at the lowest drive voltages, closely related to the drive voltage itself. Hence, feedback amplifier operational conditions are expected to be found in the following diagrams to be those along a diagonal line which, to a first approximation, is straight and passes near the origin.

The variation of amplifier characteristics with supply voltage (bias current fixed) shown in figures 6-4 a to d is in good agreement with previous data and with predictions made from waveforms found in the feedback amplifier. Although the diagrams speak largely for themselves, pertinent comments include;
Measurements Results

**FIGURE 6-4.** A & B. AMPLIFIER GAIN VERSUS SUPPLY AND DRIVE
Measurements Results

\[ x_{100} = \text{PHASE (DEG)} \]

\[ x_{10} = \text{SUPPLY (V)} \quad \text{RF DRIVE (V)} \]

\[ x_{10} = \text{SUPPLY (V)} \quad \text{RF DRIVE (V)} \]

**FIGURE 6-4.** C & D. AMPLIFIER PHASE SHIFT VERSUS SUPPLY AND DRIVE
Measurements Results

i) Contours of constant gain and of constant phase can be seen to exist over a wide signal level range, those for gain being particularly straight.

ii) Figure 6.5 shows a large scale overlay of phase and gain contours. This shows that constant phase and gain are most coincident over a relatively restricted region which lies along a line from near the origin to the point at 12 V collector supply and 4V RF drive.

iii) Also from figure 6.5 it is seen that the coincidence maximum region occurs when the amplifier is in saturation by almost 3dB, i.e. the 18dB gain and 20° phase contours are the most coincident.

iv) The bias current point chosen for these plots was that arrived at as an optimum after iterative adjustment of the feedback amplifier under dynamic (two-tone) conditions.

v) At very low signal powers, those gain and phase contours which show the greatest coincidence at medium and high powers cease to do so.

6.3.4 Input return loss versus supply

Figure 6-6 shows an early plot of input return loss. This was made using less supply voltage range than for later plots and also employed logarithmic power intervals, (which were originally chosen for increased accuracy at lower powers but proved more cumbersome in certain calculations,- see chapter 7). A ridge of increased return loss can be seen extending backwards into the projection, showing the effect of supply voltage on input impedance variation with drive level. This ridge is, unfortunately, not directly coincident with the phase/gain coincidence locus referred to in the previous section. However, since return loss is dependent upon some previously defined system impedance, the choice of amplifier conditions during the
FIGURE 6.5 OVERLAY OF GAIN AND PHASE PLOTS
input stage impedance matching procedure represents a further variable. Figures 6-7.a to d show present plots of input return loss. The input circuit tuning for the plot had previously been adjusted for a good match with the amplifier system configured in the feedback arrangement. This ensured that the measurement system impedance was met on at least one point of the locus of optimum phase-gain coincidence.

On comparison between figures 6-7.b, 6-7.d and 6-6, it can clearly be seen that there is a marked correlation between the constant, matched input return loss locus with near optimum constant gain and constant phase shift loci. This triple alignment is quite remarkable and would appear to be a previously unreported effect, except by the author.

**Figure 6-6. Early plot of return loss versus supply and drive**
Measurements Results

The input return phase angle plots show that the optimisation of match is only a transient condition, i.e. as the supply voltage is reduced (with a constant drive level), the input return loss passes through a single maximum while the return phase has a steep phase gradient with different arguments of opposing sign on each side of the loss maximum. This indicates a continuously changing input impedance as the match optimised region is traversed. Thus, depending on the adjustment of the input matching components, the input match optimisation can appear as a direct phase inversion from one side of the optimised region to the other. The implication is that input match may be used to form the basis of a further control signal.

Whether this impedance shift represents a change in the resistive component, or a change from an inductive to a capacitive reactance component, or a combination of both, is difficult to determine directly. After establishing (for calibration purposes) the return phase of a short circuit load used in place of the amplifier input, from outside the amplifier it appears as a combination of both. (It should be noticed that all of the return loss phase plots are calibrated in this manner). However, the input matching circuit adds a phase delay which complicates the issue. Further discussion on this topic is therefore left until chapter 8, on modelling.

6.4 Discussion of results

The information presented in the preceding sections shows clearly that the implementation of envelope feedback for bipolar amplifiers by power supply modulation has distinct benefits associated with it owing to particular effects in bipolar amplifiers. These effects do not appear to have been disclosed before in the literature. The simultaneous linearisation of gain,
Measurements Results

RETURN LOSS

\[ \times 10 = \text{dB} \]

A

\[ \times 10 = \text{SUPPLY (V)} \quad \text{RF DRIVE (V)} \]

B

\[ \times 10 = \text{SUPPLY (V)} \quad \text{RETURN LOSS (dB)} \]

\[ \times 10 = \text{RF DRIVE (V)} \]

FIGURE 6-7A & B. INPUT RETURN LOSS VERSUS SUPPLY AND DRIVE
Measurements Results

RETURN PHASE

\( \times 100 = \text{DEG} \)

\[ \begin{array}{ccc}
\text{RF DRIVE (V)} & 1 & 2.00 \\
\text{SUPPLY (V)} & 1 & 2.00 \\
\end{array} \]

\[ \begin{array}{ccc}
\text{RETURN PHASE (DEG)} & 1 & 2.00 \\
\text{RF DRIVE (V)} & 1 & 2.00 \\
\end{array} \]

\( \times 10 = \text{SUPPLY (V)} \)

\( \times 10 = \text{RETURN PHASE (DEG)} \)

\( \times 10 = \text{RF DRIVE (V)} \)

**Figure 6-7c & d. Input Return Angle Versus Supply and Drive**
Measurements Results

phase and input match by variation of power supply voltage also yields a fourth advantage; reducing the rail voltage also reduces DC power consumed by the amplifier. If the rail voltage can be automatically reduced using an efficient supply (such as switched mode) then the total amplifier efficiency increases dramatically at medium drive power levels. For example, in figure 6-8, at a drive level of 2V the efficiency is approximately 48% at 12V supply and approximately 78% at 4V supply. This is an important consideration for speech waveforms which normally have a peak-to-average ratio of 4:1, and which therefore implies considerable periods of time spent in the medium power region.

Even if an inefficient (e.g. series pass) modulator is used, there are still two advantages;

i) A small saving in power is made as the current consumed by the amplifier is to some extent voltage dependent (compare in figure 6-9 any point on the 12V supply axis with another point having the same drive level but at a voltage lying along the optimisation locus established in previous sections).

ii) Although this power saving is only small, a very considerable proportion of the power dissipation is shifted from the amplifier itself to the modulator, which is very useful from thermal and reliability design standpoints.

Arising from the results presented above, it follows that implementation of envelope feedback control purely by amplitude pre-distortion (i.e. with a gain controlled driver) cannot compensate for AM-PM conversion. In fact, forcing the amplifier further into saturation in order to achieve constant gain will worsen the phase modulation. Owing to a) the non-linear nature of
Measurements Results

FIGURE 6-8. EFFICIENCY VERSUS SUPPLY AND DRIVE

FIGURE 6-9. COLLECTOR CURRENT VERSUS SUPPLY AND DRIVE
the amplifier AM-PM conversion function and b) the non-linear relationship between angle modulation and sideband components (see section 4-3), the interference benefits of close-in IMP reduction by feedback are likely to be negated by the presence of the increased high order components resulting from the additional PM. By choosing the bias point correctly (i.e. for the constant phase contour condition described in section 6.3.1), this type of envelope feedback may be optimised, however saturation will occur at a relatively low peak power and the saturation distortion effects are likely to increase rapidly.

While the claim that the gain, phase and impedance effects have not been directly observed before is believed to be true, certain historical points are worth pointing out.

When AM was first applied to solid state transmitters for mobile radio purposes, the techniques used were adapted from the established thermionic valve practice of output electrode modulation e.g anode-modulation for triodes or anode-and-screen grid modulation for tetrodes. Collector modulation was found not to work directly, owing particularly to the inability of a single modulated stage to completely cut-off the RF, resulting in less than 100% modulation for negative modulation peaks. The residual carrier was described as being due to power feedthrough via the collector-base capacitance. The solution adopted for reducing this modulation distortion was to modulate several (cascaded) stages instead of only the final amplifier. Thus it would appear that at least the gain linearisation effect as isolated in this study was being used, although probably without the knowledge of the other coincident benefits.

The coincidental linearisation effects described in this
Measurements Results

chapter require explanation. Investigation of the literature on the operation and electrical modelling of RF power transistors under class-B or -C conditions reveals a very complex subject. This is therefore dealt with in a separate chapter (chapter 8).
7. RESULTS:-- FURTHER MANIPULATION

7.1 The ultimate test:-- synthesis of spectra

With any measurement system realistic assessment of total errors can be difficult. Particular difficulties exist with this system principally due to the junction temperature and therefore bias condition uncertainty under dynamic conditions (see subsection 5.4.4). The ultimate test for this system would be to re-constitute the statically measured gain and phase measurements into a two-tone spectrum. This chapter describes this process and presents the resulting spectra. Conclusions are reached which have not been presented before and which would otherwise be difficult to realise from theoretical analysis alone.

7.1.1 Requirements for spectra.

Time domain waveforms, such as a gain compressed and phase shifted two-tone envelope, have a frequency domain spectrum associated with them. The transformation between time and frequency domains may be achieved using the Fourier transform. The definition of the Fourier transform of a waveform \( f(x) \) is;

\[
S(f) = \int_{-\infty}^{\infty} s(t)e^{-j\omega t} \, dt
\]

which can alternatively be represented by the coefficients \( a_0, a_n, b_n \), defined as;

\[
a_n = \frac{1}{\pi} \int_{-\infty}^{\infty} f(t) \cos(\omega nt) \, dt \quad \quad (n=1,2,3,...)
\]
Data Processing

\[ b_n = \frac{1}{\pi} \int_{-\infty}^{\infty} f(t) \sin(wnt) \, dt \quad \text{for } n=1,2,3,\ldots \]

\[ a_0 = \frac{1}{\pi} \int_{-\infty}^{\infty} f(t) \, dt \]  

[7.2]

At each frequency \( n \), this may be viewed by electrical analogy (figure 7-1) to be two homodyne receivers simultaneously applied in phase quadrature, [1], [2], each followed by a low pass filter with infinite time constant (only passing DC) and in addition a further filter giving a direct measurement of the DC component in the original \( f(t) \).[3]

![FIGURE 7-1. FOURIER ANALYSIS ANALOGY](image)

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The evaluation of the Fourier components of real waveforms requires numerical integration techniques. From this point the limitations of sampled data systems come into play i.e. strict rules governing the minimum number of data points required for accurate analysis without ambiguity (i.e. the second Nyquist criterion). The computer evaluation of Fourier coefficients is made simple by the availability of the NAG (Numerical Algorithm Group, Oxford University) library of mathematical subroutines.

The waveform for analysis is most easily synthesised from the gain (G) and phase shift (θ) information by incorporating these into the product method for obtaining two-tones:

\[ V(t) = G(t)\cos\left(\frac{w_1 - w_2}{2}t\right)\cos\left(\frac{w_1 + w_2}{2}t + \theta(t)\right) \]

[7.3]

The large change in the frequency ratio from those used in practice might be expected, as regards distortion, to have an effect on the final spectrum. However, examination of the standard modulation equations for AM and PM reveal the modulation index and therefore sideband components to be independent of centre frequency. This is significant in reducing computer processor time.

7.1.2 Curve Fitting

The derivation of the spectrum requires the insertion of the statically measured data into equation 7.3 at regular time intervals. The resulting two-tone waveform has a peak amplitude which varies sinusoidally with time. Thus the data must be available for amplitude and phase distortion at these time intervals which are non-linear with respect to input drive power, but linear with respect to time in the final two-tone waveform.
This problem of obtaining values at the correct power points was first solved by taking measurements specifically at those points. While giving usable results, this is a limited solution as only one peak power value can be tested for any given set of measurements. A more versatile solution is to take measurements at linear drive voltage increments and then to use curve fitting techniques to interpolate the required values.

The curve fitting may be approached in many ways. In common with most statistical applications, meaningful results can only be obtained by judicious application of carefully selected algorithms. Experimentation shows that an empirical approach is the most rewarding. The original sine-plotted data points proved useful as a reference of good data.

The first decision to be taken in curve fitting concerns over which proportion of the curve the fit is to be applied at any one instant, i.e. whether an equation is found for the whole curve or just for a small section local to the required value. In this case, especially as the distortion mechanisms have not been specified in the general case, the idea of approximating the whole operational range of the amplifier by one equation appeared unwise. Hence only small sections of the characteristics are used to derive the final fit at any one particular point. The NAG library of mathematical routines has several curve fitting algorithms. The final choice of algorithm and method of implementation was made according to the following criteria;

i) The fit should, ideally, help to minimise measurement noise.

ii) Distortion mechanisms in the amplifier may lead to abrupt changes of gradient which would in turn be important contributors to the high order sidebands. These should not therefore be removed.
iii) Available fit algorithms include polynomial (general high order equation where the order is up to 1 less than the number of data points) and cubic spline (sets of cubic equations connected at user-defined points in the data, called splines, and at which the first and second derivatives of the fitted line are both zero). Both approaches employ the least-squares method.

It will be recognised that points (i) and (ii) are mutually incompatible and thus the final solution must be a compromise. A number of discussion points arise from the above;

a) Noise can be separated from high order detail by one characteristic; noise is random and only affects one individual data point in any particular fashion, whereas high order detail affects adjacent data points to some extent. If this is not true, noise and detail cannot be separated and therefore more data points are required to improve the noise performance. Systematic error and high order detail may well not be separable.

b) Use of higher power polynomials increases the ability of the fit to follow abrupt changes and therefore assists in meeting point (ii) and detracts from meeting point (i).

c) Applying a fit to a smaller number of data points increases the effect of noisy data and reduces the likelihood of a change of gradient being detected and followed.

d) Increasing the number of data points excessively reduces the sensitivity of the fit, i.e the bandwidth is effectively increased, passing greater noise power.

From the above considerations the final algorithm operated as described and shown in figure 7-2:
FIGURE 7-2. CURVE FITTING ALGORITHM

i) In the centre of the range, to establish a fitted point which lies on a measured data point $D_n$, two more data points are taken on each side of $D_n$, giving five in all. A fourth order polynomial fit is applied to this data. The fitted data point is taken as the centre point of the fit.

ii) For data points not lying on measured points, the above technique is used but the relevant point is inserted into the fit as a point between $\pm 12.5\%$ of the centre.

iii) At the extremities of the range, the most extreme five measured data points are used still, but the required data are now taken from the extremities of the fit.

The software listing of this process is given in appendix 4.
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7.1.3 Interfacing with the FFT

The NAG Fast Fourier Transform routines have to be interfaced in order to a) input the time waveform and b) recover the spectrum information output.

In the real amplifier, under two-tone conditions the ratio of RF cycles (at 29MHz) to one LF envelope cycle (tone difference of 1kHz) was typically 29000:1. To calculate the spectrum requires taking data points over one full LF period (i.e. 1mS). This would result in an excessive number of data points for calculation, wasting processor time. Thus the number of samples needs to be reduced. Bearing in mind the use of the product method for obtaining two tones, the (suppressed) carrier may be reduced in frequency, hence reducing the number of samples required and also shifting the two tone spectrum down in frequency. As the number of samples is reduced, the information pertaining to distortion reduces also. The allowable limit to this reduction is when the highest order IMP of interest or of significant amplitude (which corresponds to a particular harmonic frequency component of the modulating tone for both AM-AM and AM-PM conversions) can no longer be represented properly due to Nyquist limiting.

The spectra plotted in this chapter are made using a 360-point representation of the two-tone waveform. The representation is strictly synchronised to the carrier and modulating frequencies (hence the significance of using 360 points) in order to avoid leakage, i.e. the effective finite-bandwidth characteristics of Discrete Fourier Transform when performed on an asynchronously truncated waveform. In the course of a full two-tone waveform period, the full range of power levels is traversed four times. The gain compression and phase distortion
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data therefore need only to be calculated for 90 individual points, each point being used four times in the construction of the final waveform.

The FFT output is in the form of a Hermitian sequence i.e. the real part of the first spectral component is defined by the first entry in the (one dimensional) array while its imaginary part is contained in the last entry. The spectrum can be reconstituted from this by suitable calculation or by using a complementary NAG routine. Presenting the spectral data in the accepted histogram fashion is most easily achieved using the GINO line drawing routines.

The flow diagram for the final data processing program is shown in figure 7-3.
7.2 System Calibration

7.2.1 Initial system check

The entire measurement and data processing system as described has many possible sources of error, some of which have been discussed. The ultimate test of the system was to measure the distortion of a distortion-free amplifier. The practical solution to this was to remove the amplifier under test and reduce the high power attenuator value by the average gain of the amplifier.

Figure 7-4.a, b and c show the gain, phase and spectrum associated with an early sine-plotted scan and without curve fitting. The poor low order IMD performance was attributed to gain measurement error. Gain was defined by output voltage/ input voltage. For this scan, input voltage was measured by programming the Marconi signal generator to a voltage level calculated from a sine function equation in the PET computer and output voltage measured by the Racal level meter. The principal inaccuracy is in the voltage precision of the Marconi generator when programmed to a given resolution. As far as could be determined, the unit was designed around attenuators scaled in logarithmic intervals; the voltage mode of programming involved internal look-up tables and subsequent rounding errors (some small voltage increments appear not to change the output level at all). Also, the mechanical attenuator (which switched at 20dB increments) was not within any subsidiary control loop and caused step changes in output level and therefore in calculated gain. While these step changes cause problems in this form of data processing, the generator was not found to be out of its specification. Hence the method of obtaining input voltage for calculation purposes was changed. The most optimum solution proved to be using the second sense head of
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Voltage Gain

Phase (Deg)

Relative Input

Figure 7-4. Early System Gain, Phase and Reconstructed Spectrum
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the Racal level meter to measure a sample of the signal generator output. However, this was only after having established that each sense head has to be connected to the correct instrument port, as assigned during factory calibration and indicated by a subtle marking. The gain measurement error due to the output level error of the Marconi source now becomes a function of the change of gain with signal level in the amplifier under test, and as such was negligible.

The Racal level meter features an adjustable averaging period. The random sampling measurement technique that is employed to measure RMS voltages independent of frequency and waveform is prone to introduce noise in excess of the background noise level. This excess noise can be substantially reduced by increasing the averaging period. It was found beneficial to increase the averaging period from the default setting of 1S by a factor of four. At this setting, noise was typically ± 2 least significant digits for medium signal levels. This represents a noise voltage which is of the order of -70dB relative to the measured signal. This is considered to be adequately small.

The finally achieved limit to system performance is shown in figure 7-5, which is made using a linear plot and using curve fitting techniques. Improvements are typically of 15dB, maximising in the medium order components.

7.3 Results

7.3.1 Reconstruction of spectra

The first reconstruction of spectra from a real amplifier must still be regarded as experimental. Plate IV shows a conventionally operated linear amplifier under two different bias
FIGURE 7-5. FINALLY ACHIEVED SYSTEM PERFORMANCE
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conditions and measured on a conventional, filtering spectrum analyser. Apart from the previously observed phenomenon of different IMD slopes for the two sets of conditions, the higher bias measurement also shows another commonly observed anomaly (see section 7.3.3) of a distorted IMD slope, in this case resulting in almost equal 3rd and 5th order IMPs (plate IV, top). This anomaly represents a good test for the system. Comparison with figure 7-6.a and b, the reconstructed spectra, shows the same effect. This in particular, and also the generally good agreement of results, may be interpreted as a demonstration of satisfactory measurement system performance and also of the validity of the data processing techniques.

Direct explanations for the distorted IMD slope have not been encountered. The Sechi effect\(^49\) is a 3rd order IMD cancellation phenomenon which is optimised at a particular bias level. The cancellation condition is regarded by Sechi as an interaction between AM-PM and AM-AM distortion components, however even here, the relationship with higher order IMPs is not discussed.

7.3.2 Separation of AM-AM and AM-PM components

Owing to the use of the product method for synthesising the distorted waveform, the leaving out of phase modulation or gain compression from the waveform becomes a trivial process. The resulting spectra, however, are most revealing. In figures 7-7.a and b, the the spectra for gain compression only show the expected result that at high bias there is very little contribution to the composite spectrum especially from the medium and higher order components. Alternately at low bias, the low drive gain compression produces severe IMD. The corresponding spectra for phase modulation only (figures 7-8.a and b) show that
FIGURE 7-6. RECONSTRUCTED SPECTRA FOR CONVENTIONAL AMPLIFIER
(AS PLATE IV)
Two spectra of the test amplifier under conventional conditions of fixed bias and supply. The commonly found phenomenon of equal 3rd and 5th order products forms a useful test for calibrating the spectral reconstruction program.
Data processing

the converse is true, although not to the same extent when considering the very high order components.

Examination of figures 7-7.a and 7-8.a shows clearly that at this particular drive level, at higher bias, the 3rd order IMD is almost completely due to gain compression while the 5th order IMD is mainly (but not so uniquely) generated from the phase modulation. Thus separation of the complete spectrum reveals the heavy influence of phase modulation.

7.3.3 Effects of power level

The previous section described the anomalous effect of a distorted IMD slope. It was stated that the figures presented were those as measured at a particular power level. Thus there is interest in establishing the effect of power level on the IMD spectrum generally. Figures 7-9.a and b show two tone spectrum variation with power level. These were achieved by the previously described method of curve-fitting to linearly measured data points, the change of power level for each 'slice' of spectrum being achieved by purely computational means.

It should be noted that cancellation effects exist such that individual IMPs are minimised at particular power levels. Also, higher order IMPs enjoy the same cancellations but at higher drive levels, which results in these figures as valleys of IMD minima, extending at some angle to the direction of increasing drive level. It should further be noted that the IMD does not reduce with power level in the same manner as would be expected of small signal amplifiers (when distortion levels are relatively much reduced even under worst case conditions) i.e. not according to the law as indicated in figure 1.2 that an n dB reduction of peak power level results in a 3n dB reduction in IMP power for
FIGURE 7-7. SPECTRA FOR GAIN COMPRESSION ONLY
FIGURE 7-8. SPECTRA FOR PHASE MODULATION ONLY
FIGURE 7-9. RECONSTRUCTED TWO-TONE SPECTRA VERSUS DRIVE
Data Processing

3rd order components, 5n for a 5th and so on. By examination of figures 7-10.a - d, where gain and phase effects with power level have been separated on measurements performed at the significant bias currents of 70mA and 400mA (as discussed in section 6.3.1), this unexpected result can be seen to be due to either gain or phase non-linearity dependent on the bias level. From figures 6-2.b and d, these are either or both severely non-linear even towards low power levels.

7.4 Discussion of results

From an extensive study of the literature carried out by the author there has been found little detailed discussion of the effects of bias point on the performance of normally operated solid state class-B tuned amplifiers, nor of the significance and effects of AM-PM conversion. The data processing carried out in this chapter culminating in the diagrams 7-10.a to d show beyond question that these two points are linked and that bias conditions critically control the nature of the distortion. It should be noted that without separating the phase and gain distortion components, this change is not easily deduced from the spectrum alone. As adjustment of practical amplifiers yields complex and confusing changes in the IMD spectrum, it is standard practice to adjust SSB transmitter amplifiers only for low order product amplitude performance. However, this clearly cannot separate or quantify the proportion of contributions from the phase shift or gain compression modes of IMD production.

Distortion spectra for solid state amplifiers show an increased proportion of high order products (alternatively, lower slope) when compared with thermionic amplifiers operated under similar low order distortion conditions.\textsuperscript{93} Initial consideration of this might lead to the supposition that this is purely due to
FIGURE 7-10A & B. GAIN COMPRESSION SPECTRA VERSUS DRIVE
FIGURE 7-10C & D. PHASE MODULATION SPECTRA VERSUS DRIVE
Data Processing

the AM-PM conversion effects which are known to be present in the solid state amplifiers, the non-linear nature of PM sidebands relative to the modulation (as mentioned in section 4.3.1) being the principal cause for the lower slope. However, a clearer description can now be given as follows (see figures 6.2-a to d):

i) At low quiescent currents (approx. 1% of maximum signal current) the distortion is principally due to the non-linear gain-versus-power relationship which steepens towards lower powers, subsequent interaction with the fast crossover transitions present in two-tone waveforms resulting in high order IMD production.

ii) At higher quiescent currents (approx 10% of maximum signal current) the gain becomes much more constant and the AM-PM conversion distortion dominates. However, the simpler nature of the AM-PM conversion law results in a greater slope and therefore reduced high order IMPs.

iii) Operation above the linear point involves more similar contributions from both gain compression and phase shifting. The saturation point of such amplifiers used for linear applications requires definition. For 12V devices it is common practice for device manufacturers to recommend operation to be maintained less than 60% of the CW rated power. Examination of figure 6-2.d (phase shift versus drive and bias) shows that even along the contours of constant phase which are linear with increasing drive (70mA bias), the phase shift clearly becomes severely non-linear above a certain power level, which is well below the power saturation point (the manufacturers rated output power for CW applications).
8. MODELLING THE AMPLIFIER

To this point in the thesis much time has been spent on measurements of a real amplifier in which a BJT shows particular effects of interest, notably those of obeying previously undiscovered facts of constant gain, phase and input impedance. These subsequently allow new applications of such amplifiers which have also been shown. The true nature of these simultaneous linearisation effects in terms of device operation have remained unexplored. This chapter sets out to begin along this path.

8.1 Approaches to Modelling

Modelling the amplifier may take one of several very different forms. The following descriptions represent distinct categories and it should be noted that a practical modelling solution may well be a hybrid of these.

8.1.1 Systematic modelling

Systematic modelling deals with the development of a model composed of synthetic elements such as perfect amplifiers, phase shifters and variable attenuators, whose values vary with drive power and supply rail. When working from measured data, this is likely to lead to an accurate model when used for simulating the amplifier as part of a larger system. Since the model only contains those elements necessary to model the effects, it is likely to need least computational power and therefore offer the highest speed. This will be useful when implementing certain mathematical techniques for non-linear distortion modelling (i.e. Volterra series, see appendix 2). This might subsequently prove
Modelling

to be the best route for analytical techniques to follow towards the calculation of distortion performance of the amplifier operated under some particular feedback regime, or alternatively when modelling to optimise the feedback regime for ultimate distortion performance. However, this type of approach does not take into account the device physics in any way and hence there will not necessarily be any connection between the operation of the synthetic elements and any recognisable equivalents in the real amplifier. Also, on the premise that only carrier fundamental amplitude and phase performance are used in the synthesis of the model elements, the unipolar nature of real amplifiers could not possibly be modelled, thus the likelihood of generating elements similar to established BJT model elements is still further reduced.

8.1.2 Mathematical modelling

The amplifier may be treated purely as a black box with no deliberate intent to relate the amplifier properties to circuit elements. Successful characterisation of a greater number of the variables (e.g. terminating impedances, carrier frequency, power level, signal bandwidth etc) is likely to result, with the increased chance of predicting performance in the general case.

8.1.3 Large signal BJT equivalent modelling

The previous two methods have been abstract to a greater or lesser degree. An alternative approach is to model the amplifier (i.e. matching components, supply impedances and BJT) by non-linear circuit means i.e. all passive amplifier components are represented as such while the BJT is modelled using a non-linear model of lesser or greater complexity and either in mathematical or non-linear component form. This method is the most flexible
and offers the greatest chance of accurate modelling in order to establish electrical equivalents for physical properties of the semiconductor device and thus the chance of leading to useful information (especially from the point of view of device manufacturers) pertaining to the transistor characteristics. The increased computational complexity will result in much greater processor time requirements, and also a greater degree of effort required for accurate synthesis of elements than for the previous method. However, this approach is favoured here owing to its pragmatism.

8.2 Established BJT models

Electrical circuit models of active electronic devices have been developed historically very soon after the invention of the device in question. This occurs in order that such devices may be incorporated into conventional (linear) circuitry with calculable performance and also as a ready means of describing the physical behaviour of the device in a language familiar to the engineering fraternity. This is true whether the device is a thermionic valve (where the output electrode is modelled as a current source programmed by the input electrode voltage) or a rotating machine (where the winding resistance and inductance, and back-EMF generators are the important components). The accuracy and therefore complexity of the model is usually determined by operational requirements, with simplicity always kept in mind. The ubiquitous availability of digital computers in modern day laboratory situations has tended to allow considerably greater-complexity to enter into such models without exceeding allowable computational time limits than would have been true beforehand.

Within the category of transistor amplifier modelling already defined, BJT models take many forms, depending on the
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application and the user. For general transistor applications, models which show the non-linear effects associated with the solid state physics of such devices must be used i.e. containing representations of:

i) Exponential voltage-current relationships.

ii) Capacitance effects from separated charge regions and from charge storage effects.

iii) Bulk resistance effects.

The most familiar models to the electronic circuit designer are those for small signal applications and which do not contain signal dependent non-linearities, e.g. the hybrid-pi model. For all signals within the working range of the circuit, the model values are considered constant and therefore the model can be used for the calculation of gain, frequency response, etc. by the application of linear circuit theory. For high frequency applications scattering (or s-) parameters are more commonly used, where the model reduces to a two-port "black-box", the device ports then being characterised by the voltage reflection and coupling co-efficients of power waves arriving at the ports when connected in a 50-ohm system. Such models allow sufficiently accurate engineering calculations to be made, but mask the further appreciation of device physics. It is not uncommon, however, for non-linear models to be developed forms of the hybrid-pi model.

For BJTs intended for power applications at HF and VHF (both CW and linear), it is current practice to specify the impedance which the device input presents (and which is conjugately matched to), and the circuit admittance (or, sometimes, impedance)
Modelling

required by the device as an output load. Examination of manufacturers data reveals that such information is generally established in applications laboratories by purely empirical means\textsuperscript{95}, modelling the device beyond this being largely avoided. Papers are regularly found in the literature which intend finally to replace such techniques with more accurate modelling,\textsuperscript{96,97,98,99} however their regularity implies that the perfect solution has not yet been found and meanwhile the empirical methods persist. Indeed, in one reference\textsuperscript{95}, engineers working in this field for two decades still denounce the successful use of theoretically derived models.

The difficulties associated with this topic can easily be demonstrated. Figure 8-1 shows the major components of a typical tuned amplifier at VHF, working at a power level of a few tens of watts i.e. such that the input and output impedances are low. A typical input impedance (as taken from a manufacturers data sheet) could be 2 ohms (taken to be resistive for this example).

\textbf{FIGURE 8-1. TYPICAL TUNED AMPLIFIER}
Modelling

Input matching is achieved by employing the current magnifying effect of the tuned circuit $Q$, the transistor input impedance being the series resistance element of the tuned circuit. Thus using the approximation that the parallel impedance of a tuned circuit is $Q^2$ times the series resistance, a $Q$ of 5 leads to an input impedance of 50-ohms. Now, as the voltage across the tuned circuit is a sinewave, the current in the tuned circuit is also a sinewave. Hence the base current of the transistor is a sinewave. This is a major anomaly as transistor theory indicates that the device should become high impedance for negative currents (accepting that the device is working in class-B), thus open circuiting the tuned circuit once in each half cycle. Examination of typical base waveforms reveals only small voltage swings in comparison with the voltage across the input tuned circuit, indicating indeed that the base input current is essentially a sinewave. This anomaly exists also for the output tuned circuit.

To this point the grossly simplified models and equivalents considered are those promoted and used by industrial and commercial engineers. Solid-state device designers and academic researchers have tended to approach modelling from a different angle, i.e. extensions or modifications of more mathematically based models developed in the 1950's, and derived chiefly from solid-state physics considerations. The gap in styles of approach just described is quite marked, with the literature showing a notable absence of cross references between the two schools. With the advent of the integrated circuit design engineer, however, (where the circuit engineer becomes his own device engineer) the gap has begun to be bridged. The bridging is made feasible by the use of complex computerised modelling programs which allow the designer to accurately model circuits before fabrication. Such programs saw considerable development in the early 1970's, and titles of some of these also imply significant interest at that
Modelling

time in the effects of radiation on complex circuitry\textsuperscript{100}.

The most important early large signal mathematical model of a BJT (and which, with extensions, forms the basis of the majority of commercial modelling programs) is the Ebers-Moll model of 1954\textsuperscript{101}. In this the chief objective was to model the DC terminal behaviour of the device in the three regions of operation i.e. cut-off, amplification, and saturation. The equivalent circuit, consisting of current generators (called "alpha" generators) and diodes with exponential voltage-current relationships, is shown in figure 8-2. The Ebers-Moll equations are of the form:

\[ I_E = a_{11}(e^{V_E/V_T} - 1) + a_{12}(e^{V_E/V_T} - 1) \]  
\[ I_C = a_{21}(e^{V_E/V_T} - 1) + a_{22}(e^{V_E/V_T} - 1) \]

\textbf{FIGURE 8-2. EBERS-MOLL EQUIVALENT CIRCUIT}
Modelling

It can be seen that the model actually consists of two transistors, one inverted over the other. These enable the four-quadrants of operation to be correctly represented. In practice, the inverted transistor has approximately 10% of the gain of the normal transistor.

![Graphs showing Ebers-Moll and Gummel-Poon DC characteristics](image)

**FIGURE 8-3. EBERS-MOLL AND GUMMEL-POON DC CHARACTERISTICS**

The Ebers-Moll model has certain failings, which are compensated for by extensions and later models. The Gummel-Poon model of 1970 is generally regarded as modelling most effects of importance. The important extra characteristics include;

1) Early effect\(^\text{102}\). Figure 8-3.a shows I-V characteristics as represented by the Ebers-Moll model. In practice, however, the current generator impedance represented by the transistor has a finite conductance and results in plots such as 8-3.b. This was identified by Early as being due to the effects of base width modulation as a result of the collector depletion (space-charge) region widening,
resulting in increased current gain at higher collector voltages. The effect can be characterised most easily by extrapolating the output characteristics back to the point of cutting the \( V_{CE} \)-axis. This point is common for a wide range of base currents and can be represented by the crossing point, called the Early voltage.

ii) Bulk resistivity resulting in finite collector and emitter resistances.

iii) Kirk (base push-out) effect \(^{103}\). At high collector currents the \( f_T \) and \( h_{fe} \) of BJTs is observed to reduce. This is owing to high current densities causing significant electric fields in the collector region, resulting in widening of the base region \(^{104},105,106,107\). In the circuit representation of the Gummel-Poon model this appears as abruptly increasing emitter diffusion capacitance above certain combinations of collector current and voltage.

iv) Low-bias dependence of \( h_{fe} \). The emitter part of the base-current (\( I_{be} \)) is modelled by two diodes in parallel, one ideal and the other with an ideality factor of greater than 1.

8.3 Implementing Spice

Currently several circuit modelling programs are commercially available which involve more or less accurate implementations of established transistor models. For this study, an attempt has been made to use SPICE (Simulation Program with Integrated Circuit Emphasis), which has been developed at University of California (Berkeley). The model used is an extension of the Gummel-Poon model, the use of default values for
certain parameters resulting in a simplification back to the Ebers-Moll model.

8.3.1 Introduction To SPICE

SPICE models may consist of passive components (resistors, capacitors, inductors and transmission lines), voltage sources, and current sources. The sources may be dependent, also capacitors and inductors may be non-linear functions of voltage and current respectively. Models are supplied (but with user variable parameters) for the diode, BJT and FET. Circuit optimisation is not a feature of this program and component value sensitivity checks are only available under limited conditions.

There are three major modes of operation. These are;

i) DC analysis. All inductors become short circuits and all capacitors open circuits. This can be used independently and is automatically run before a transient analysis in order to establish the initial conditions, and also before an AC small signal analysis to establish any non-linear model characteristics to be used in the analysis.

ii) AC small signal analysis. After establishing the AC small signal circuit values, frequency transfer characteristics may be calculated. As the circuit is using the small signal values which are (assumed to be) linear, amplitude distortion cannot be taken into account using this method.

iii) Transient analysis. Using the previously established initial conditions, repeated transient solutions may be taken. By reducing the iteration period sufficiently, non-linear circuits driven with sinusoidal sources may be accurately
Modelling

modelled. Under these conditions model parameters vary with the conditions as found at the start of each iteration.

Clearly, the transient analysis mode is the only one suitable for power amplifier modelling.

8.3.2 Constructing the model:— Passive components

The real circuit is sensitive to tuning conditions. As the circuit was deliberately built to be similar to the application note design, with inductor values arrived at by dimensional specification, initial values for modelling the input and output circuits was achieved by direct measurement of component values when dismantled from the amplifier. Subsequent trimming of the input components was possible by substituting a synthetic network for the input impedance of the transistor with values taken from the manufacturers application data.

The trimming process consists of measurement of the input VSWR. Figure 8-4 shows the circuit of the input. The signal source has its output impedance controlled by resistor Rs. VSWR is measured as the voltage between the effective amplifier input terminal (node 1) and a high value potential divider Rbr1 and Rbr2, and this may be seen as a bridge circuit. The high value of Rbr1 & 2 leads to negligible extra current drawn from the signal source. A match is achieved when the voltage at the bridge point is nulled out. As the matching circuit employs its Q to achieve the impedance transformation, the network is of limited bandwidth and therefore of finite risetime. Several cycles of the 30MHz driving waveform must pass before the true value of the input impedance is reached. The transient analysis is performed with approximately 30 iterations per cycle. Thus several hundred iterations are required for such a measurement.
Although the real amplifier was measured at 29MHz, this was changed for the model by a small percentage to 30MHz in order that the Fourier transform routine could successfully operate. Apart from the need for slight retuning, it is not anticipated that any effects in the real circuit are sufficiently frequency sensitive to cause any significant errors.

The use of empirical techniques for achieving an impedance match with such a trivial network may appear odd. However, this was also intended as a skill acquisition procedure in anticipation of the needs of the complete amplifier. Other techniques for establishing the input impedance of the amplifier directly are not considered applicable as it is expected that the complex nature of the amplifier is effected by the impedances presented by the matching networks to the input ports at harmonic frequencies.

The power supplies have been assumed to be zero impedance voltage sources. In the real amplifier great care was taken to
ensure that this was true for all intents and purposes by the liberal application of wideband decoupling networks, consisting of carefully selected and paralleled capacitors and also series resistors for resonance damping.

Observation of amplifier waveforms show little effects due to temperature other than those of bias and thus these have not been included.

Waveforms of the real amplifier are distorted by the bandwidth limitations of the oscilloscope. For the model, therefore, waveforms have been deliberately distorted also, by the inclusion of filter networks with risetimes and circuit loading characteristics similar to those of the real instrument.

8.3.3 Device parameters

SPICE allows the user to specify 40 parameters. The full listing for the BLW60 and the method used for obtaining them is described in appendix 1. Those parameters found to be of greatest importance are discussed below:

1) Inter-electrode capacitances. Solid state device capacitances are of great importance in the design of BJT power amplifiers. From section 8.2, the fact that the base terminal input current is substantially sinusoidal (while the intrinsic transistor must clearly be working over severely non-linear portions of the device characteristics) implies that the larger portion of the terminal current is flowing into the device capacitances.

SPICE requires the capacitances to be specified at zero bias and with an exponential or junction grading factor. Capacitance measurements were taken at 1MHz and using 100mV
Modelling

Measurement of the capacitances in isolation is not possible, and only composite values can be measured (figure 8-5). Analysis to establish a general formula for calculating the individual values results in awkward non-linear equations. Thus the simplest way is to use computerised iteration techniques. The collector and emitter junction grading co-efficients were established from plots of capacitance versus reverse bias voltage.

\[
\begin{align*}
C_1 &= CX + \frac{CYCZ}{CY + CZ} \\
C_2 &= CY + \frac{CXCZ}{CX + CZ} \\
C_3 &= CZ + \frac{CXCY}{CX + CY}
\end{align*}
\]

**FIGURE 8-5. DEVICE CAPACITANCE MEASUREMENT**

ii) Substrate junction. The majority of modern power transistors have built in emitter ballast resistors. This is principally because such units consist of many individual devices in a large array and thus the increased risk of thermal runaway needs reducing. (The chief mode of breakdown is called second breakdown\textsuperscript{108} and is ultimately due to an excessive
Modelling

instantaneous current-voltage product in an individual transistor element, resulting from a localised thermal runaway effect known as "hot-spotting"). In the fabrication of the resistors, however, a new diode junction is created between collector and emitter, which is reverse biased under normal working conditions. The presence of this junction is not common knowledge amongst RF engineers.

Effects which arise include an extra capacitance between collector and emitter of approximately 20% of the zero bias collector-base junction capacitance, and most significantly, severe inhibition of normal device operation in the reverse active region. This poses problems when one attempts to determine the reverse characteristics (which are important in the modelling of charge storage effects).

SPICE allows for a junction between the collector and the substrate (intended for integrated circuit work) which may be connected as an extra node, or to ground by default.

iii) DC parameters. Beta, Early voltage, high current Beta roll-off and low current Beta effects (the latter principally controlled by leakage parameters) can all be established from DC measurements, those for the forward direction with good accuracy. However, the reverse active region figures can be estimated only by extrapolation from between the origin and the ballast resistor-collector junction cut-in voltage.

iv) Parasitic elements. Although bulk resistivity effects are available in SPICE, internal inductances are not. These have to be included as parasitic elements outside the device. The most significant is emitter inductance, which although only of 1nH, develops more than 1 Volt of potential difference due to the RF emitter current. This voltage is, of course,
Modelling

directly added to any measurement of base voltage.

These elements are of further significance owing to resonance effects with other external reactances at high frequencies, which become shock excited by frequency components arising from the mechanisms of normal transistor action.

8.4 Results

DC parameters for the model were obtained by a combination of direct measurements, use of manufacturers published data specific to the device used and estimations derived from other published works on computer modelling of similar devices. These other works were also found very useful guidance towards most of the other parameters.

Initial attempts to model the amplifier were performed on a simple tuned amplifier arrangement, the intention being to optimise only later the effects due to more elusive components such as parasitic inductances (which were originally anticipated to be only small, especially when considering that this device is being modelled at only 20% of its maximum rated frequency). However, it was quickly found that these other elements are quite crucial to the effective modelling of large signal amplifiers, due (a) to the very low impedances and large currents involved in such amplifiers and (b) to the fact that these type of devices are being used to the limits of their capabilities in order to achieve the claimed gains, powers and efficiencies.

A new approach was therefore needed to establish order of magnitude values for the device. A transient method was developed using square-wave base current pulse injection, carried out on the device in common emitter mode with a short circuit (AC)
Modelling

collector load (fig.8-6). The resulting base voltage transient is shown in figure 8-7. The use of a grounded collector removes the Miller-effect of the collector-base junction capacitance, which otherwise severely affects the HF response to the current transient.

**FIGURE 8-6. CIRCUIT OF BASE IMPULSE MEASUREMENTS**

The initial rising slope in the base voltage waveform is due to the superposition of the emitter voltage, which itself is developed from the emitter current passing through the emitter parasitic (lead) inductance, figure 8-8. This was established purely by replication on the SPICE model. (It should be emphasized that the emitter tabs were connected to the ground plane with the minimum effective lead length which could practically be achieved).
Modelling

![Graph of External Vbe Impulse Response](image)

**FIGURE 8-7. BASE VOLTAGE TRANSIENT RESPONSE**

![Graph of Emitter Inductance Impulse Response](image)

**FIGURE 8-8. Emitter Inductance Transient Response**

After this method had been taken to its useful limit, further adjustment of the model parameters by comparison with the real device could only be made with the model embedded in the matching circuitry. The problems of iterating a now complex system were approached by attempting to imitate on the computer model as many measurement techniques as were practicable on the real amplifier. These included:

i) Input reflection measurements (amplitude and phase).
Modelling

ii) Base terminal voltage waveform.
iii) Collector terminal voltage waveform.
iv) Output load voltage waveform.
v) Collector current DC component.

Measurements of high impedance points of tuned circuits were specifically avoided in order to minimise the risk of confusion by spurious loading effects, which would have been difficult to measure and therefore to imitate accurately.

Certain parameters are easily measurable on the computer model but cannot be made at all on the real amplifier. The best examples of this are all RF current measurements. Observing such waveforms in the model leads to considerably improved perspectives over several aspects of amplifier operation. To a limited extent, this still applies even when the model is not working accurately, e.g. that the base terminal current is nearly sinusoidal is true for all but the most severe forms of modelling error. Other measurable parameters exclusive to the model include those voltage waveforms inside the parasitic lead inductances.

The components and model optimisation iteration procedure proved to be somewhat circuitous and involved occasional sudden changes as new elements were either introduced for the first time or were brought to a suddenly more realistic value. The development of the values will not be further discussed here, however, instead the final model waveforms will be discussed in relation to those of the real amplifier and the two sets of component values. The final model circuit is shown on figure 8-9.
Modelling

FIGURE 8-9. SPICE AMPLIFIER CIRCUIT
8.4.2 External Amplifier and model waveforms.

a) Input waveforms

The input reflection bridge waveform (figure 8-10) takes several cycles to reduce to its balanced condition. In comparison with the same waveform resulting from matching the input network into a dummy input impedance, the settling time is increased. This implies reduced bandwidth and stems from the interaction between input and output matching circuits. Substantial reverse coupling is also implied.

The bridge null waveform also shows back-injection of harmonics generated within the amplifier. This is in line with the inability to achieve 1:1 SWR when using wideband SWR meters (see section 4.3.2).
Modelling

Base terminal input current is nearly sinusoidal and peaks at 1A for higher drive levels. The base bias current, however, only reaches approximately 10% of this. The implication is that the majority of the base terminal current passes through the device capacitances without resulting in transistor action. Although the RF base current cannot be measured on the real amplifier, the voltage across the input matching circuit inductor can be deduced to be nearly sinusoidal and hence the stated current waveform follows, i.e. for the impedance levels involved, the input matching circuit appears to the transistor input as a good approximation to a current source.

Base terminal voltage is non-sinusoidal even at low currents. As drive level increases the general shape is distorted by the successive addition of two nearly sinusoidal kinks (figure 8-11 and plate V). These result from RF emitter current being reflected in the emitter impedance and superimposed on the base waveform. The waveform of the emitter current is described later.

b) Output waveforms.

The voltage across the load resistance RL is of good sinusoidal shape at low power levels. However some distortion can just be seen at the highest power levels.

The collector current and voltage waveshapes are somewhat drive dependent, being sinusoidal at low drive levels. Saturation effects in the voltage waveforms of the real amplifier are initially difficult to explain, i.e. the apparent saturation voltage of approximately 3V, with a central inflection (plate V). The model did not show the same effects at first but required the addition of parasitic collector inductance and
RF Drive: 1.5V RMS

RF Drive: 3V RMS

RF Drive: 4V RMS

RF Drive: 5V RMS

PLATE V

Amplifier Waveforms Varying with Amplitude

Each oscillograph shows the 29MHz collector waveform (10V/div, upper trace, centre line zero) and base waveform (1V/div, lower trace, zero line marked) at different drive levels (horizontal scale 100μS/div). The 4V drive case is that normally found under optimised envelope feedback conditions. Collector supply was 12V.
FIGURE 8-11A. COLLECTOR AND BASE WAVEFORMS VERSUS DRIVE
FIGURE 8-11B. COLLECTOR AND BASE WAVEFORMS VERSUS DRIVE
Modelling

resistance plus parasitic lead inductance in the collector capacitor, \( C_{\text{out}} \).

8.4.3 Internal Waveforms

One of the great advantages of the SPICE model is that voltage and currents may be measured at any point in the circuit without the practical problems of disturbing the circuit or requiring careful engineering to avoid stray impedances from affecting the measurements. Hence the voltages present at the intrinsic transistor nodes may be observed, yielding surprisingly conventional results.

a) Emitter-ground voltage (figure 8-12). The voltage at the emitter device node (71) approximates to the differential of the emitter current, due to the parasitic emitter inductance

\[
\begin{align*}
\text{FIGURE 8-12. INTERNAL EMITTER-GROUND VOLTAGE}
\end{align*}
\]
Modelling

L_{pe}. This voltage is superimposed on the base voltage (figure 8-11) and can be seen to be the major part of the base voltage and therefore to have a large influence on the input impedance, i.e. output to input coupling (or feedback) occurs in these elements.

b) Base-emitter voltage (figure 8-13). Despite the base to ground voltage of figure 8-11 showing considerable base voltage excursions, intrinsic device measurements on the model show that in fact the voltage swings only up to 0.8V and down to around 0V. Experimentation with capacitance values in the device model showed that the negative excursions especially are limited principally by the presence of the non-linear collector-base capacitance with the (approximately) phase inverted voltage at the collector, and secondarily by the base-emitter capacitance.

c) Collector-emitter voltage (figure 8-14). This waveform is shown at a medium drive level and clearly shows a fascinating saturation characteristic to be present inside the parasitic elements. As has already been stated, in a classical class-B amplifier the current waveform is assumed to be half sinusoidal. However in the type of amplifiers considered here (i.e. with low output impedance feeding an upward-transforming matching network) the current can be explained and measured to be sinusoidal. Instead it is the intrinsic device voltage waveform which is half sinusoidal, i.e. these amplifiers represent the dual of conventional thermionic amplifiers. This also implies that, as far as engineering the output network is concerned, the amplifier output impedance should be considered as a voltage source instead of a current source.
Modelling

**Figure 8-13. Intrinsic Base-Emitter Voltage**

**Figure 8-14. Intrinsic Collector-Emitter Voltage**
Modelling

8.4.4 Explanation of external waveform non-linearities.

Several features of the voltage waveforms found around the test amplifier require explanation:

1) Collector voltage inflection. While the magnitude of the sinusoidal component of the collector current across the collector and emitter parasitic impedances would be expected to show some voltage drop, this cannot explain the extent of the inflection found in practice. Also, the inflection is related (in time) too closely for coincidence to the kink in the collector current waveform (figure 8-16). Hence it is proposed that when the intrinsic collector-emitter junction saturates two things occur; Firstly high frequency components are produced by the rapid change of slope in the waveform and secondly the collector-emitter junction becomes a very low impedance. The output matching circuit becomes modified under these conditions (due to the collector choke being largely shorted out) and a previously unobserved tuned circuit becomes excited at a resonance near the third harmonic. This new resonant circuit has as its inductance the emitter and collector terminal inductances, $L_{ep}$ & $L_{cp}$, plus the collector capacitor lead inductance, $L_{oc}$. This capacitor is the tuned circuit resonating capacitor and Q is limited by the loss resistance of the capacitance plus the bulk and ballast resistances within the transistor. To test this theory, (a) the resonant frequency was computed as below and (b) experimentally it was shown that reducing the capacitor loss resistance to zero resulted in increased ringing, see figure 8-15.

ii) Base voltage kinks at high drive levels. This derives from the same mechanism as (i) and is the emitter inductance component of the oscillation superimposed onto the true base waveform.
Modelling

**FIGURE 8-15. EFFECT OF COLLECTOR CAPACITOR LOSS RESISTANCE**

**FIGURE 8-16. COLLECTOR CURRENT WAVEFORM**
Modelling

iii) Non-sinusoidal base voltage at low drive. Even at the lowest drive levels the base terminal voltage waveform is non-sinusoidal (see figure 8-11.a and plate-V). This results from the nearly sinusoidal collector, and therefore emitter current being differentiated by the emitter inductance to form the (now phase-shifted) emitter voltage, which has superimposed on it the non-linear base voltage (an exponentially modified sinusoid, see figure 8-13).

8.4.5 Multi-dimensional measurements.

By embedding the SPICE modelling program into a Command Programming Language (CPL) program, it proved possible to make multi-dimensional plots of gain and phase shift, much as the direct measurements plots were made in chapter 6. Figures 8-17.a to d show gain and phase versus bias and drive while figures 8-18.a to d show the same against supply and drive. The features of constant gain and phase with supply and drive are clearly present, as are bias currents which result in constant gain and constant phase over a limited drive range and with fixed supply. There are, however, two important differences.

i) For variation of bias with fixed supply, the regions of constant gain and constant phase are coincident. This is not true in the real amplifier, see figure 6-2.

ii) Phase shift characteristics are similar in all cases, except for the magnitude of the shifts which are considerably reduced in the model.

Input impedance against supply and drive is shown in figure 8-19. The linearisation characteristic is clearly observable for this also.
Modelling

FIGURE 8.17A & B. SPICE AMPLIFIER GAIN VERSUS BIAS AND DRIVE
Figure 8-17C & D. SPICE Amplifier Phase versus Bias and Drive
FIGURE 8-18A & B. SPICE AMPLIFIER GAIN VERSUS SUPPLY AND DRIVE
Modelling

**Figure 8-18C & D.** SPICE AMPLIFIER GAIN VERSUS BIAS AND DRIVE
Modelling

RETURN LOSS
$x10 = dB$

$x10 = SUPPLY (V)$

$3.50$

$3.00$

$2.50$

$2.00$

$1.50$

$1.00$

$0.50$

$1.2$

$1.0$

$0.8$

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Figure 8-19c & d. SPICE input return loss versus supply and drive
8.5 Conclusions

By modelling using an established transistor model (the Gummel-Poon model), and by accurately modelling other circuit components using inductive, resistive and capacitive elements where necessary, good correlation has been obtained over a wide dynamic range for waveforms, power gain and input impedance characteristics between an iterative transient analysis circuit model and a real, BJT large signal power amplifier.

It has been stated many times in the literature that RF power transistor amplifier circuit analysis proves impossibly difficult and it is beyond the scope of this thesis to alter this. However, taking established facts as described in the literature and arising out of the author's work, certain arguments may be formulated that may lead nearer to the correct analysis of such amplifiers.

Facts already established from published works and from the modelling performed to date are:-

i) The base terminal current is nearly sinusoidal and is considerably in excess of the intrinsic device base current that would be expected for the typical collector terminal currents found.

ii) The collector voltage waveform becomes an increasingly truncated sinusoid as power level increases.

iv) From plate VI, the region of simultaneous linearisation occurs when, despite power level changing, the collector voltage saturation angle remains constant, which is ensured by the choice of collector voltage supply.
PLATE VI

Envelope Feedback Amplifier Waveforms

These two oscillographs of collector (upper trace, centre line zero) and base voltages (lower trace, zero line marked) were taken from an envelope feedback amplifier at two levels of drive which differed by 6dB. The oscilloscope calibration was also switched by this amount (2:1) between exposures and a shift of base voltage zero was required for the low power case. The overall similarity of the two exposures is remarkable.
Modelling

From these it may be concluded that:

i) Gain compression in this type of amplifier is brought on by collector voltage saturation.

ii) Because of the region of constant gain by supply control, the gain compression mechanism is independent of high current Beta or \( f_t \) effects.

iii) Saturation angle is the controlling function.

iv) Limiting the output voltage swing affects the feedback via \( C_{bc} \). This causes the change in the input impedance which is also related to saturation angle.

No direct explanation of the phase shift effects is available, especially as in this respect the model fails to completely agree with the real system. However, plots of the real amplifier show a phase characteristic with a steep central slope, which is very similar to the classical phase shift versus frequency characteristic for a simple tuned circuit. This implies an effect in the amplifier which alters the tuning of one or both of the matching networks with drive power and saturation angle.

Considerable experimentation with model parameters has failed to isolate one with a large phase sensitivity. Those parameters which do yield a greater effect also reduce the gain to unrealistic values. Precise values for many such parameters are generally unobtainable in a direct form either from measurements or from discussion with device manufacturers, and hence the need for empirical adjustment. Further work would most beneficially involve close communication with solid-state specialists.
9. CONCLUSIONS

9.1 Review of objectives

The original study brief was to investigate Ultra-Linear amplifiers as intended for use in the power amplification stages of radio transmitters. Initial investigation quickly confirmed that although in essence all transmitters exhibit the same problems of distortion and DC-RF power conversion inefficiency, bandwidth and absolute frequency considerations dictate large differences of approach to be suitable for each case.

Non-linearity in amplifiers results in spurious products. These can be located at harmonic frequencies of the wanted signal and also, by the process of intermodulation when two or more signal frequencies are simultaneously present, at all frequencies related arithmetically to the original signal frequencies and their harmonics. (If a waveform being amplified has a time-varying amplitude, it can be analysed as a multiple frequency waveform and consequently will suffer intermodulation distortion.)

Such products, when generated by Radio transmitters, usually cause interference to other spectrum users and it is for this reason that such products have to be removed. In the case of a spot frequency transmitter (whether it be radio communications, broadcasting or television), harmonics do not cause insoluble problems as they can be removed by filtering. However, some of the IMP's are located at frequencies immediately adjacent to the wanted transmission and therefore cannot be filtered out in any practical situation. In the case of T.V. transmitters the most
Conclusions

troublesome product is actually within the transmission band itself and causes degradation of the received picture.

In an age of rising dependence upon radio communications both for civil and military applications the interference problem becomes less tolerable and more intense as;

i) Spectrum congestion increases resulting in closer adjacent channel frequency spacings.

ii) The spatial density of transmitting stations increases resulting in extended wanted signal to unwanted signal ratios likely to be experienced at the receiving site.

iii) Channel allocation plans become increasingly unable to optimise (i) in conjunction with (ii).

iv) Portable transmitting equipment becomes more common, resulting in increased demand for highly power efficient transmitters. (In conventional transmitters, linearity and therefore interference performance worsen with increasing efficiency and in the words of one senior engineer "the less batteries a soldier has to carry, the more bullets he can carry".

v) Fixed transmitting equipment (especially for broadcasting applications) encounter increased running costs particularly owing to the rising price of energy.

Accepting that the products exist, one solution in the case of a single transmitter is to use modulation techniques which do not involve time-varying amplitude components. Frequency Modulation is one example of this. However, this has problems of
Conclusions

its own, particularly poor performance at low received signal-to-noise ratios and wider bandwidth required than for Amplitude Modulation techniques. This last point is of growing importance with the increasing world usage of radio communications generally. It was decreed by the American Federal Communications Commission in 1956 that SSB should be used for all HF point-to-point links and it now seems not unlikely that a move to SSB at VHF will be necessary (unless low bandwidth constant amplitude digital techniques are developed first). It is accepted that there is no longer any more spectrum available for radio communications as techniques have been developed for using all those frequencies with usable propagation characteristics. Thus it is now a matter of managing with what is currently available or developing more spectrally efficient techniques. Hence as SSB is the most spectrally efficient speech mode in common use, highly linear amplifiers are required.

For HF and VHF communications systems it is becoming increasingly common at multi-user sites to employ low power signal sources and one combining (or Multiplex) power amplifier. In these amplifiers modulation technique does not have the same bearing as for the single transmitter amplifier as by their very nature there is more than one signal present and thus intermodulation can occur. Also it is not uncommon (especially at HF) for multiplex amplifiers to cover more than one octave and hence harmonic and even order intermodulation products now fall in-band and cannot be filtered out. Thus wideband techniques must be employed. There follows a brief historical outline of how the above objectives have been tackled in the past.

9.1.1 Historical review.

Examination of the literature reveals several methods for
Conclusions

reducing IMD in power amplifiers and these are described more fully in chapters 2 and 3. The earliest of these were either intended for AM transmitter applications or in telephone line FDM repeater amplifiers. Most of the methods appearing in the current literature were first invented in essence during the 'twenties and 'thirties, but were either too difficult to implement or have otherwise been re-invented without knowledge of the first appearance. In many cases, especially when mathematical analysis is carried out, the principles can be traced back still further to analogies in mechanical systems or in chemical process engineering. Basically four techniques have emerged:

1) Feedforward. Conceived in 1923 by H.S.Black and patented in 1929, potentially capable of very great improvements in amplifier performance but difficult to implement with operational stability, this idea has not caught on in commercial applications. It has received attention on a regular basis through the decades and it would appear to have been re-patented on at least one occasion. Most success so far reported is by H.Seidel of Bell Labs, who achieved 42dB improvement over the band 0.5-20MHz in a telephone FDM amplifier and 50dB improvement in a microwave Travelling Wave Tube application using two feedforward loops. These results were published in the seventies. After another quiet period the US Naval Research Labs have recently published data on an HF Multiplex amplifier employing feedforward which achieves -85dBc for two-tones at 80% of peak power. The largest problem is to maintain the cancellation condition. The NRL system uses some adaptive control to achieve this, demonstrating a path for the future of wideband amplifiers.

2) Feedback. Also conceived by Black but some years after feedforward, this technique has been widely used for many decades in its direct form at audio and low radio frequencies. However as
Conclusions

it is a post-distortion correction system, significant delay through the amplifier ultimately limits its usefulness. At microwaves where amplifier delays may be several cycles, it is clearly not applicable at all. Even at HF and VHF its use is not always possible because of accumulated delays and unpredictable phase shifts in tuned circuits and non-resistive loads.

Traditionally feedback has been applied by sampling either output voltage or output current using resistive techniques. This results in a worsening noise figure with increasing feedback and hence there is a new trend in "loss-less" feedback, which yields wider dynamic range by using broadband transformer directional coupler techniques.

Feedback need not be applied directly to the signal at the signal frequency and ingenious applications of feedback to other parameters of amplifier performance have been developed. In 1941, Terman and Buss used the detected envelope of an AM signal as the feedback parameter in a linear amplifier, the output of the feedback network controlling the amplifier by grid modulation.

3) Parameter separation. In 1952 Kahn published details of an experimental system for modifying existing AM transmitters for SSB use. The low power SSB signal is separated into amplitude and phase components and each path is amplified separately, being combined in a conventional anode modulator at the high power end. Since then Petrovic has developed this into the so called Polar-Loop system where much the same idea is used except that the phase component is translated from I.F. by the use of phase-locked-loop techniques and also the envelope signal is kept low distortion by the use of feedback derived from synchronous detectors. There are difficulties with this system due to the very wide bandwidths being required in the phase path. For this
Conclusions

reason Petrovic has recently developed a new quadrature feedback system called Cartesian-Loop, which shows considerable promise.

4) Pre-distortion. This technique has generally only been widely used in TV transmitters, where gross distortion is needed if any efficiency whatever is to be obtained. Because of the high peak-to-average power ratio for TV signals, efficiencies of less than 5% are not uncommon for older equipments currently in service.

9.2 Experimental work and results.

This project has included a large proportion of practical work. Of this, the important experiments pertain to measurements of the variation of gain, phase shift and input impedance with power supply and bias conditions for tuned BJT power amplifiers at HF and VHF.

It is widely known that gain compression causes intermodulation distortion and thus it is common for transistor data to include plots of output power versus input power. It is also widely known that AM-PM conversion results in IMD. However there is only a minimal amount of qualitative or quantitative information regarding AM-PM in this type of amplifier available in the published literature. The results obtained in this thesis show that the degree of phase shift and therefore AM-to-PM conversion is highly significant. Moreover, the amplifier phase shift is affected by supply rail voltage and bias conditions.

Measurements performed on this type of amplifier have shown that under conditions of fixed supply:-

1) Quiescent bias current can be chosen to minimise gain
Conclusions

compression or AM-PM conversion (but not both) over a limited range of power upwards from zero.

ii) Bias pumping cannot be used to any significant advantage for combating gain compression and AM-PM conversion simultaneously.

iii) Lower bias currents produce reduced low-order IMPs but the exponential characteristic of the resulting crossover-distortion leads to increased high order IMPs. Hence (a) higher bias currents are to be preferred and (b) specification of IMD by low order measurements alone is misleading.

Also, under conditions of fixed bias it is possible to dynamically alter the supply rail voltage such that:-

i) The relative phase shift at all drive powers is kept to a very small figure.

ii) Gain compression is minimised

iii) Input impedance is simultaneously linearised.

iv) Efficiency is maintained to much lower drive levels due to the correct supply impedance (voltage to current ratio) being maintained.

An amplifier was constructed which operated over the simultaneous linearisation regions described above, the control voltage being derived by conventional (peak detector) envelope feedback. That the amplifier can also be operated over the same region with the control voltage derived either from measurements of phase or input impedance is currently the subject of a multiple patent application in Europe, Japan and the USA being undertaken on behalf of the author.

Performances achieved during this research project were 50%
Conclusions

DC-RF conversion efficiency (using a switched mode supply modulator) with 3rd order IMP's approaching 40dB. The results proved repeatable on a day-to-day basis and were implemented at 30 and 150 MHz. There are difficulties in establishing the correct feedback conditions especially for the very low power regions, high order products being particularly vulnerable to incorrect settings.

Processing of statically measured gain and phase-shift data led to the re-construction of two-tone spectra. These implied the following conclusions:-

i) Static measurements of fundamental frequency gain and phase shift are all that is required to perform two-tone distortion analysis of an amplifier, even though the amplifier contains severe non-linearities and may be experiencing very large internal circulating currents at harmonic frequencies. From this, IMD prediction from non-linear circuit modelling using computer techniques only need accurately reproduce these characteristics (i.e. gain and phase-shift) over the required power range.

ii) In conventional amplifiers the contributions from gain compression and AM-PM conversion are dramatically varied by bias conditions

iii) The frequency symmetry found in the re-constructed spectra disagrees with the established belief\(^\text{113}\) that spectral asymmetry for two-tone distortion products is due to the simultaneous presence of both gain compression and AM-PM conversion (which is a well known result for conventional AM). It was discovered and checked by data processing that asymmetry under two-tone conditions can, in fact, be due to
Conclusions

phase shifts at the difference frequency of the two tones, and most likely can be the result of poor decoupling in power supplies.

iv) Conventional class-AB power amplifier IMD performance does not improve with reducing peak power level in the same calculable manner as with untuned class-A amplifiers. This is due to the exponential gain compression at the crossover region (low bias conditions) or to AM-PM conversion effects (high bias conditions). Complex interactions between the gain and phase effects lead to cusps in the IMD curves of individual products. Cusps occur at different power levels for different products and are generally observed to progress to higher order products as the peak-power level increases.

Non-linear circuit modelling using the Gummel-Poon model in a pseudo-linear transient analysis program has been attempted, enabling deeper understanding of the nature of operation of solid-state tuned power amplifiers. BJTs intended for RF power applications have historically been considered to be beyond analysis and it is common practice for manufacturers to only state large signal impedances for such devices. Waveforms found within the model at the intrinsic device nodes show that the operation of such amplifiers most closely resembles the dual of the conventional thermionic class-B amplifier in that it is the collector voltage which is half-sinusoidal and not output current, as is classically the case. Hence there is an anomaly:— Should the device output be modelled as a current source (as represented in the classic Ebers-Moll model) or as a voltage source (which is suggested by the implication of the dual of the thermionic class-B case)?
Conclusions

After achieving satisfactory correlation in the model between voltage waveforms, power gain and phase shift effects, multi-dimensional plots were performed similar to those previously taken on the real amplifier. The results may be summarised as follows:-

i) The regions of constant gain, phase and input impedance with adjustment of supply and drive were also found to exist in the model.

ii) The regions in the conventional (fixed supply) amplifier of constant gain and phase at particular quiescent bias currents were found. However these were (a) not at the same bias currents as in the real amplifier and (b) occurred at exactly the same current, unlike the real amplifier.

iii) Phase shift effects in the model were very similar, except reduced by a factor of three over those found in the real amplifier. Time has prohibited any further analysis of this last point.

iv) The regions of simultaneous linearisation coincide with constant intrinsic collector saturation angle.

9.3 Suggestions for further work

i) To continue analysis with SPICE modelling to improve accuracy especially for phase shift effects, finally processing data into IMD spectra by employing techniques as used in this thesis. Subsequently use the developed model to investigate the possibility of non-linear predictions of other amplifier designs, using computer analysis to save bench time.
ii) To continue to optimise envelope feedback amplifiers, possibly by (a) improving envelope detectors, and (b) using combinations of envelope, phase and input impedance information to derive the control signal.

iii) To consider the use of amplifiers operated in this manner within other established forms of distortion reduction e.g a Cartesian Loop transmitter, or a feedforward system.

9.4 Overall Conclusion

Bipolar tuned linear power amplifiers at HF and VHF create IMD by gain compression, AM-PM conversion and by presenting a dynamic load impedance to the driving stage. The contribution of each of these mechanisms is not widely appreciated or understood, and is affected by amplifier conditions, especially quiescent bias current.

However, such amplifiers may be operated in a new linearised mode, as suggested by the author, by the application of feedback. The feedback signal may be derived from envelope, phase shift or input impedance information and used to control the collector supply voltage.

The most notable resulting improvements are:

i) Considerable reduction of medium order IMD products.

ii) Improvement of DC-RF power efficiency for the amplifier, especially at lower powers.

iii) Reduction of heat dissipated in the amplifier, even if the
Conclusions

power saving cannot be achieved in the supply modulator stage.

iv) Simultaneous linearisation of gain, phase-shift and input impedance. The linearisation of input impedance results in reduced back-injection of IMD products.
APPENDIX 1

SPICE Amplifier Input Listing

BLW60 30MHZ POWER AMPLIFIER
VCC 9 0 12
VIBASE 20 5
IBIAS 0 40 5E-2
VRF 13 0 SIN(0 14 30MEG)
EBIAS 41 0 40 0 1.0
VEBASE 41 4
DBIAS 40 0 D1 23.15
.MODEL D1 D RS 0.641
RBR1 13 130 1000
RBR2 130 0 1000
CBSCOPE1 21 0 5E-13
RSSCOPE1 3 0 50
RSSCOPE2 3 21 1000
CCSCOPE1 31 0 1E-12
RCSCOPE1 8 31 500
LIN 2 3 150NH
RS 13 1 50
LC1 1 100 2NH
C1IN 100 2 26PF
C2IN 2 200 161PF
LC2 200 0 2NH
RBIAS 3 4 22
LBIAS 3 4 2E-7
LC 8 9 60NH
CC 8 81 450PF
RCC 81 80 7.0E-3
Appendix 1

LCC 80 0 2E-9
LOUT 8 10 110NH
C1OUT 10 0 10PF
C2OUT 10 11 700PF
C3OUT 11 0 400PF
RL 11 0 50
Q1 6 5 7 73 Q1 2000
VSCAPM 73 72
RE1 7 72 0.010
RE2 72 71 0.010
LE 71 70 0.8E-9
LCM 1 8 1.5NH
RCI 60 61 0.15
MODE XQ1 NPN BF 50 BR 5 VAF 60 CJE 345E-15 CJC 142E-15 MJC 0.33
+KXJC 0.6 TF 2.0E-10 RB 100 RE 5 CJS 62.5E-15 VAR 20 IS 1.0E-16
+ITF 5.0E-3 XTF 2.0 PTF 180 TR 6E-9 VTF 10 RC 1 MJS 0.5 MJE 0.3
+IRB 2.5E-3 IKF 5.0E-3 NR 1.0 NF 1.0 IKR 3.0E-4 ISC 1.0E-13
+FC 0.5 ISE 1.0E-13
LBP 3 20 1E-9
CBP 3 0 2PF
VECM 70 0
VCCM 60 6
.TRAN 1000PS 1000NS 0 1000PS
.OPTION LIMPTS=20002 ITL5=20001 TNOTM=27
.FOUR 30MEG V(11) V(13)
* PRINT TRAN V(5,7) V(21) V(72) I(VIBASE) V(31) I(VCCM) V(1,130)
* PRINT TRAN V(1) V(2) V(3) V(1,130) V(5) V(6) V(7) V(20)
* PRINT TRAN V(5,7) V(8,0) V(6,0) V(6,7) I(VCCM) V(71,70)
* PRINT TRAN I(VSCAPM) V(5,0) V(7) V(8,60) V(6,5) I(VIBASE)
* PRINT TRAN V(5,7) V(6,7) I(VCCM) V(6,5) I(VSCAPM) V(1,130)
* PRINT TRAN V(1,130) V(31) V(21) V(7) V(5,7) V(6,5) I(VCCM)
.END
Non-linear systems have always been of concern in communications and have historically defied accurate analysis. It is normal in many cases that the non-linearities are only weak e.g. distortion effects in class-A line repeater amplifiers or in radio receiver front-end stages. Control engineers also face this problem, especially in the event (under commercial pressures) of having to engineer systems from available or lower cost components (with associated linearity limitations), instead of using more expensive components in genuinely linear regions. In the event of strong non-linearities, the problems of analysis become considerably more complex and the mathematics for this has been, and is currently, mainly a research topic amongst theorists.

A2.1 The Power Series approach

The classical method is to treat the non-linear element as an instantaneous non-linearity, and to approximate this with a power series, as equation (A2.1).

\[ y(t) = ax(t) + bx(t)^2 + cx(t)^3 \]  

An important stipulation is that the non-linearity must not have as part of its output any term dependent upon previous conditions i.e. that it is "memoryless". Thus an amplifier whose ports are connected to systems with memory, i.e non-resistive, with energy storing elements such as resonant filters, may be
Appendix 2

correctly analysed only if the ports are non-interacting. Consider a field effect transistor amplifier with negligible inter-electrode capacitance, whose output is a current source controlled only by the gate-source voltage. This description fits the term memoryless. However, consider further the event of a close spaced two-tone waveform and also that the source-ground connection includes an inadequately bypassed biasing resistor (figure A2-1). The system now contains memory; the largely square-law characteristic of the FET will rectify the instantaneous magnitude of the input waveform and the resulting LF component will appear across the source impedance, this now constituting a frequency conscious feedback path. The result is that in practice, all terms will be modulated by the LF term arising from the rectification process (which is principally a 2nd-order IMP), and will develop sidebands accordingly, the higher order terms being distorted with greater complexity.

![Diagram](image)

**FIGURE A2-1. DE-COUPLED FET AMPLIFIER**

The classical model calculates the output waveform, which is a complex function of time, assuming superposition to be valid. Also, as there is no facility for entering parameters with time dependence, it is assumed that the effective controlling voltage
Appendix 2

is known at all times and is constant. Clearly, if the system has memory, this is not true and in practice the instantaneous value of power series expansions vary during the waveform, yielding erroneous results for power series calculations.

A2.2 The Volterra Series representation

One method to overcome this is by using Volterra series representation. It is well known that for a linear system with memory, (i.e. resistors, capacitors and inductors) the output time response to an input waveform \( x(t) \), is the convolution integral

\[
y_1(t) = \int_{0}^{t} c_1(t - T)x(T)dT
\]

This is effectively stating mathematically that the response of a linear system to a continuous function is equal to the superposition of multiple system responses to a train of closely spaced impulses which are scaled according to the original continuous function. (A.2) may be re-written in the transform domain as

\[
Y_1(s) = C_1(s)X(s)
\]

For a non-linear system also containing a term in \( x^2(t) \), this may be expanded to form the double convolution integral

\[
y_2(t) = \int_{0}^{t} \int_{0}^{t} c_2(t - T_1, t - T_2) \prod_{i=1}^{2} x(T_i)dT_1dT_2
\]
The output depends on the past values of the input; the above expression involves a product of the input with itself, thus representing a quadratic system. $c_2(t_{-1}, t_{-2})$ is known as the second degree Volterra kernel. For representation in the transform domain, this now requires a two-dimensional Laplace transform

$$Y_2(s_1, s_2) = C_2(s_1, s_2) \prod_{i=1}^{2} X(s_i) \quad [A.5]$$

This term contains frequency sensitive information with respect to two frequencies, and the kernel $C(s_1, s_2)$ is interpreted physically as the voltage transfer ratio for second order terms (i.e. $(f_a + f_b)$ in the case of a two frequency input). This method of representation may be continued ad infinitum, an infinite sum of integrals of this form being called the Volterra series.

It is important to observe that this method calculates all possible combinations of frequencies, including multiple combinations of one frequency with itself. Thus, for example, gain compression is arrived at by terms from second and higher order interactions which subtract from the first order (or linear) component. Each route to an individual component of a real frequency component in the output waveform is referred to as a frequency mix (and should be distinguished from the conventional concept of frequency mixing).

The Volterra or non-linear transfer function method has seen considerable attention in recent years\cite{45, 114, 115, 116, 117}. However, it has a number of drawbacks;
Appendix 2

1) The method requires that the system be only weakly non-linear.

ii) Computation of all but the lowest order products becomes unwieldy and very costly in processor time.

iii) For successful implementation, the kernels require accurate specification.

Thus for power amplifier applications (where hard nonlinearities are involved) the method is unsuitable. It has enjoyed favourable publicity, however, in applications concerning telephone line repeater amplifiers\textsuperscript{118} and communications receivers\textsuperscript{119}. It is significant that in ideal circumstances both of these only suffer from low order distortion problems.

A2-3 Miscellaneous techniques.

Applications for intermodulation calculations are abundant in communications engineering, especially for frequency planning at multi-user sites and in transmitter/receiver design. Because of this many little known methods have been published which in general enable quicker or simpler calculation of the most offending products and their approximate amplitudes \textsuperscript{120,121,122,123,124,125,126,127}.

A2-4 Chaotic non-linear differential equations.

A recent piece of non-linear theory which may have future relevance concerns systems describable by second order, non-linear differential equations. It has been shown\textsuperscript{128,129} that under certain conditions, (i.e. with a large driving function)
such mathematical systems show random and therefore unpredictable behaviour, termed chaos\textsuperscript{130, 131}. Such behaviour showing wideband noise components is also found in RF engineering practice (and is often referred to as instability), commonly in association with severely mis-terminated tuned amplifiers, or in circuits or systems which include unintentional signal loops which subsequently oscillate.
C  ********************************************************************** 3-D FOURIER ANALYSIS PROGRAM  **

IMPLICIT DOUBLE PRECISION(A-H,O-Z)
INTEGER I, IFAIL, NF, N, J, KPLUS1, IFAV, IFIT, I3D, J3D,
  * IGS, IGAP, ICTR, JCTR, NA, IQ, K, ICONT
REAL Y, V, R, T, VW, W, AZ, HT
DOUBLE PRECISION XFIT, XOUT, S, PI, XPUT, FWR
DIMENSION YAV(400), X(400), Z(400), ARRAY(6, 6), XAV(400),
  * WORK1(3, 5), WORK2(2, 6), ARRA(6), WAV(400), SI(4), GAV(400),
  * YPV(400), ARRPy(6, 6), ARRFX(6), W(1050), AZ(30, 26)
DIMENSION VW(400), T(400), V(400), EAV(400), E(400), R(400)
DIMENSION VI(400, 2), VO(400, 2), PH(400, 2), SC(400, 2)
DIMENSION RI(400, 2), PR(400, 2)
CHARACTER STRING*30
CHARACTER FILEA*10
CHARACTER F1*5, F2*5, F11*5, F0*1
PI=DAC0S(-1.0D00)
F0=CHAR(1)
F1=F0//'@'
F2=F0//'A'
F11=F0//'J'
KPLUS1=5
IFAV=0
IFIT=0
PRINT*, 'READING FILE'
FILEA='PUCCA6'
OPEN (2, FILE=FILEA)
READ (2, '(A)') STRING
READ (2, '(A)') STRING
Appendix 3

READ (2,*) (VI(I,1),VO(I,1),PH(I,1),SC(I,1),BC(I,1),RI(I,1),
* PR(I,1),N,I=1,101)
READ (2,*) (VI(I,2),VO(I,2),PH(I,2),SC(I,2),BC(I,2),RI(I,2),
* PR(I,2),N,I=1,101)
CLOSE (2)

1 PRINT*, 'POWER LEVEL BELOW 5V (DB)'
READ *, IPWR
PWR=-14.5-IPWR
J3D=1
PRINT*, '500 OR 50 MA (2,1)'
READ *, IIQ
PRINT*, 'SPECTRUM OR GRAPH'
READ *, IGS
FGD=1.
FPD=1.
PRINT*, 'GAIN (1) OR PHASE (2) DELETED'
READ *, IGPD
IF (IGPD.EQ.1) FGD=0.
IF (IGPD.EQ.2) FPD=0.
PRINT*, 'NUMBER OF INTERVALS'
READ *, NA
J3D=1

C ************ LOADING REQUISITE NO. OF POINTS INTO ARRAY **

IGAP=100/NA
DO 10 I=1,NA+1
ICTR=(I-1)*IGAP
IF (I.EQ.1) ICTR=1
IF (I.EQ.101) ICTR=100
G(I)=VO(ICTR,IIQ)/(VI(ICTR,IIQ)*14.00)
E(I)=PH(ICTR,IIQ)
T(I)=SINGL(E(I))

193
CONTINUE

C ************************** LOADING GROUP AND AVERAGEING
2 JCTR=2
   K=0
20 DO 21 J=1,5
   XAV(J)=J
   YAV(J)=G(J+(JCTR-2))
   YPV(J)=E(J+(JCTR-2))
   WAV(J)=1
21 CONTINUE

C **************************************** FIT CURVE TO GROUP **
CALL EO2ADF(5,KPLUS1,6,XAV,YAV,WAV,WORK1,WORK2,ARRAY,SI,IFAV)
CALL EO2ADF(5,KPLUS1,6,XAV,YPV,WAV,WORK1,WORK2,ARRPY,SI,IFAV)
DO 30 J=1,KPLUS1
   ARRAX(J)=ARRAY(KPLUS1,J)
   ARRPX(J)=ARRPY(KPLUS1,J)
30 CONTINUE

C ************************** ESTABLISH XFIT

40 K=K+1
   XFIT=DSIN(DBLE(K)*PI/180.)*DBLE(NA)*(10**((PWR/10)))
   XFIT=(XFIT-DBLE(JCTR))/2
   IF (XFIT.LE.0.5) GOTO 50
   IF (JCTR.GE.(NA-2)) GOTO 50
   K=K-1
   JCTR=JCTR+1
   GOTO 20

C ************************** FIT TO CURVE

50 CALL EO2AEF(KPLUS1,ARRAX,XFIT,XOUT,IFIT)
CALL E02AEF(KPLUS1, ARRPX, XFIT, XPUT, IFIT)
GAV(K) = XOUT
EAV(K) = XPUT
IF (K.LE.89) GOTO 40
PRINT*, FWR, XFIT, J3D, JCTR

C *********************** GENERATING 360 POINT DATA

DO 60 J=1,360
K=J-1-180
IF (K.LT.0) K=-K
K=K-90
IF (K.LT.0) K=-K
IF (K.EQ.0) K=1
S=DBLE(J)
R(J)=REAL(J)
PI=DCOS(-1.0D00)
Y=DCOS(1.0*(S-1.0)*2.*PI/360)
* *DCOS((9.0*(S-1.0)*20.*PI/360)+FPD*EAV(K)*PI/180.)
X(J)=Y*(GAV(K)*FGD+(1.-FGD))
60 CONTINUE
IF (IGS.EQ.1) GOTO 70
C LISTING OF DATA + STOP
CALL T4010
PRINT*, F11
CALL PICCLE
CALL GRAF(R, T, (NA+1), 0)
CALL DEVEND
PRINT*, F1
GOTO 1
STOP
70 IFAIL=0
CALL C06EAF(X, 360, IFAIL)
Appendix 3

DO 80 J=1,180,1
Z(J)=(X(361-J)**2+X(J+1)**2)**0.5
VW(J)=SNGL((10.*LOG10(Z(J)**2))-13.90)
IF (VW(J).LT.-70.) VW(J)=-70
80 CONTINUE
DO 81 I=87,93,2
PRINT*,VW(I)
81 CONTINUE

C ************ FIT INTO ARRAY IN READINESS FOR 3-D PROJECTION

DO 85 I3D=-25,25,2
AZ(J3D,(I3D+27)/2)=VW(I3D+90)
85 CONTINUE
J3D=J3D+1
PWR=PWR+0.5
IF (J3D.LT.31) GOTO 2

C *************** 3-D PROJECTION

PRINT*,F11
CALL T4010
CALL HEIRAT(0.75)
CALL ISOPRJ(30,-15.,0.,26,-25.,25.,AZ,0,1050,W)
CALL DEVEND
PRINT*,F1
PRINT*,"ENOUGH OR MORE"
READ*,ICONT
IF (ICONT.EQ.1) GOTO 1
END
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