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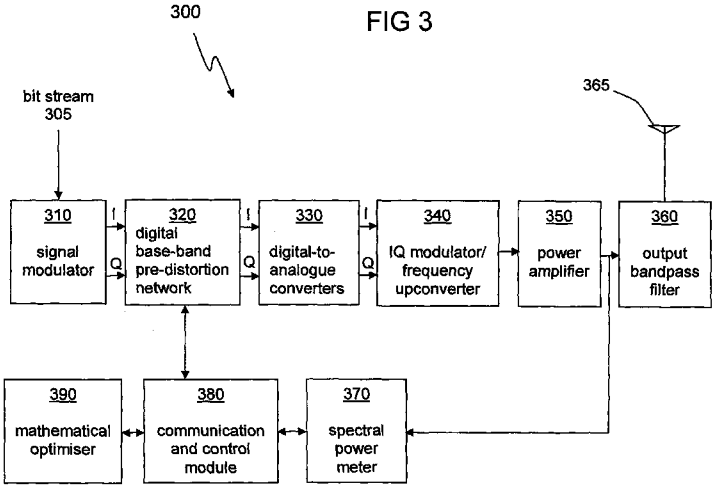
FIG 3



(57) Abstract: A method for linearising a multi-carrier radio frequency transmitter or a multi-user CDMA radio frequency transmitter enables improved linearisation. The method includes measuring a function of out-of-band signal power in the frequency domain at an output of the radio frequency transmitter. A digital base-band pre-distortion, performed by a digital base-band pre-distortion network, is then applied to the radio frequency transmitter according to the measured function of the out-of-band signal power.

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TITLE

A METHOD AND SYSTEM FOR LINEARISING A RADIO FREQUENCY TRANSMITTER

FIELD OF THE INVENTION

This invention relates to a method and system for linearising a radio frequency transmitter and in particular a multi-carrier radio frequency transmitter or a multi-user CDMA radio frequency transmitter.

BACKGROUND TO THE INVENTION

Radio frequency (RF) transmitters, such as mobile phone base station transmitters and digital broadcast transmitters, exhibit non-linear transfer characteristics due to Field Effect Transistor (FET) semiconductor devices and Class AB push-pull amplifier operation. FIG 1 shows a graph 100 of a transmission 110 in the form of a Digital Video Broadcasting — Terrestrial (DVB-T) transmission output from a typical RF transmitter, as is known in the art. As a consequence of the non-linear transfer characteristics of the transmitter, the transmission 110 is distorted causing spectral regrowth. Spectral regrowth is classified as either co-channel distortion 111 (distortion within the allocated transmission channel 120), upper adjacent channel distortion 112 (distortion in upper adjacent transmission channels 130) and lower adjacent channel distortion 113 (distortion in lower adjacent transmission channels 140).

The role of the transmitter's output bandpass filter is to remove Adjacent Channel Distortion (ACD). FIG 2 shows a graph 200 of a typical frequency response of the output bandpass filter. However, as the output bandpass filter has a finite roll off and attenuation 210 some ACD is still transmitted compared with an ideal frequency response 220. The ACD ultimately acts as interference to other users of the RF spectrum. In order to control this form of interference, regulatory authorities impose strict spectral emission limits in the form of a spectral mask. As shown in FIG 1, the transmission 110 exceeds a regulatory spectral mask 150. A desired output 160 from the RF transmitter, which is below the spectral mask 150, is shown by a dashed line.

Although a receiver may filter out ACD received from the intended transmission due to its greater bandpass filter selectivity at the intermediate frequency (IF), the receiver cannot filter out co-channel distortion. As a result, co-channel distortion interferes with the intended broadcast, resulting in symbol constellation warping/spreading (and therefore symbol detection errors) and an increased Bit Error Rate (BER).

In contrast with Amplitude Modulated (AM) and Frequency Modulated (FM) signals which have constant envelopes, multi-carrier Orthogonal Frequency Division Multiplexing (OFDM) signals and multi-user Code Division Multiple Access (CDMA) signals have non-constant envelopes and a higher Peak to Average Power Ratio (PAPR). As a result, for the same average transmitted output power, multi-carrier OFDM and multi-user CDMA signals demand greater transmitter linearity as large signal peaks drive the transmitter into regions of greater non-linearity causing greater distortion and spectral regrowth.

One method of improving the linearity of a transmitter is to perform Output Back Off (OBO). OBO involves backing off the input signal power such that the transmitter output is operating in a near linear region. This is undesirable however as the transmitter's efficiency reduces.

Several techniques exist for improving the linearity of a transmitter without performing OBO. These include active biasing, feed-forward, negative feedback, LINC, analogue RF pre-distortion and digital base-band pre-distortion. A clear distinction should be made between the latter two pre-distortion techniques.

Digital base-band pre-distortion involves inserting a non-linear discrete-time/digital network directly at the output of the transmitter signal modulator at base-band. This network is referred to as the digital base-band pre-distortion network. The digital base-band pre-distortion network's non-linear transfer characteristic is designed to be the inverse non-linear transfer characteristic of all transmitter components following the signal modulator, thereby creating an overall linear cascade. Note the use of the term "network" here refers to any system which processes its input to produce an output.

Analogue RF pre-distortion on the other hand involves inserting a non-linear continuous-time/analogue (as opposed to digital) network directly at the input of the transmitter power amplifier at RF (as opposed to base-band). This network is referred to as the analogue RF pre-distortion network. The analogue RF pre-distortion network's non-linear transfer characteristic is designed to be the inverse non-linear transfer characteristic of just the power amplifier alone. An example of an analogue RF pre-distortion technique is disclosed in a paper by Rey ("Adaptive Polar Work-Function Pre-distortion" IEEE Transactions on Microwave Theory and Techniques, VOL. 47, NO. 6, JUNE 1999) where the pre-distortion is applied according to a function of the out-of-band signal power in the frequency domain.

However, digital base-band pre-distortion has several major advantages over analogue RF pre-distortion including better cost effectiveness, reconfigurability, superior design of the non-linear transfer characteristic, improved adaption and the ability to linearise the entire transmitter, not just the power amplifier.

Some existing digital base-band pre-distortion techniques are described in the following publications:

1) Hyun Woo Kang, Yong Soo Cho, and Dae Hee Youn, IEEE Transactions on Communications, Vol. 47, No. 4, April 1999, *"On Compensating Nonlinear Distortions of an OFDM System Using an Efficient Adaptive Predistorter"*;

2) Jian Li and Jacek Ilow, Proceedings of the 3rd Annual Communication Networks and Services Research Conference (CNSR '05), May 2005, *"A*

Least-Squares Volterra Predistorter for Compensation of Non-linear Effects with Memory in OFDM Transmitters";

3) European Patent Publication, EP 1 203 445 B1;

4) US Patent No. 5,900,778;

5) Nima Safari, Joar Petter Tanem, and Terje Roste, IEEE Transactions on Microwave Theory and Techniques, Vol. 54, No. 6, June 2006, *"A Block Based Predistortion for High Power-Amplifier Linearization"*;

6) Ezio Biglieri, Sergio Barberis, and Maurizio Catena, IEEE Journal on Selected Areas In Communications, Vol. 6, No. 1, Jan. 1988, *"Analysis and Compensation of Nonlinearities In Digital Transmission Systems"*;

7) Qian Yeqing, Li Qi, and Yao Tianren, Proceedings of the 2003 IEEE International Conference on Acoustics, Speech and Signal Processing (ICASSP '03), Vol. 2, April 2003, *"Analysis of Different Predistortion Structures and Efficient Least-Square Adaptive Algorithms"*; and

8) P.L. Gilabert, G. Montoro, and E. Bertran, Microwave Conference Proceedings 2005 (APMC 2005), Asia-Pacific Conference Proceedings, Vol. 2, Dec. 2005, *"On the Wiener and Hammerstein Models For Power Amplifier Predistortion"*

Three main problems with existing digital base-band pre-distortion techniques are as follows:

1. The measure of transmitter output nonlinearity used to drive the pre-distortion network adaption algorithm is a **time domain** mean squared error between the signal modulator output and transmitter output. This measure requires a full feedback path between the transmitter output and signal modulator output which incorporates signal delay and gain compensation, RF to base-band frequency translation and analogue-to-digital conversion. In practice however the time domain signal is not a pure measure of the transmitter output non-linearity. The time domain signal is a measure of all the imperfections of the transmitter

plus all the imperfections of the feedback path. These imperfections include:

- Signal Delay and Gain Compensation Error (Feedback Path)
- Analogue-to-Digital Converter Distortion (Feedback Path)
- Carrier/Local Oscillator Leakage (Transmitter & Feedback Path)
- Phase Noise (Transmitter & Feedback Path)
- Linear Distortion (Transmitter & Feedback Path)

Thus the actual measure can be corrupted / impure and does not truly represent the transmitter output nonlinearity. This leads to sub-optimal performance in driving the pre-distortion network adaption algorithm and hence overall suboptimal linearisation performance.

2. Pre-distortion network coefficients are derived by mathematically inverting a behavioural model of the transmitter. This behavioural transmitter model is obtained via system identification techniques which in general require known test signals (possessing desirable characteristics) to be injected into the transmitter. In practice, this means that any time the pre-distortion network coefficients are to be updated (in order to track changes in the transmitter's non-linear transfer characteristic occurring over time); the transmitter must be taken off-air so that the known test signals can be injected (in place of the normal broadcast signal). Given an off-air transmitter is highly undesirable for the transmitter operators, the pre-distortion network coefficients are rarely updated, despite needing to be, and hence the transmitter is poorly linearised for the majority of its operational life.
3. Another problem exists with deriving pre-distortion network coefficients via mathematical inversion of a behavioural transmitter model. As signal modulation bandwidth increases, so too does the difficulty in accurately modelling the transmitter's frequency

dependent characteristics and memory effects. As a result, for the wideband signals used in modern transmission formats, larger inaccuracies exist in the transmitter model and hence the mathematically inverted pre-distortion network. This ultimately leads to poor linearisation performance.

With respect to addressing problem 1 above, a paper by Stapleton, S.P. and Cavers, J.K, (*"A New Technique For Adaptation of Linearizing Predistorters*", 41st IEEE Vehicular Technology Conference 1991, Gateway to the Future Technology in Motion, 19-22 May 1991) includes an analytical investigation into a method for linearising power amplifiers in which a frequency domain (as opposed to time domain) measure of output nonlinearity is used to drive the pre-distortion network adaption algorithm. This frequency domain measure does not require a full feedback path and is hence error free. However due to the simplistic and memory-less amplifier and pre-distortion network models used, this technique is only suitable for linearising amplifiers for narrowband, constant envelope transmissions and is thus unsuitable for modern multi-carrier OFDM (DVB-T, DAB, 4th generation mobile OFDMA) and multi-user CDMA (3rd generation mobile wideband CDMA (WCDMA)) transmissions which are wideband and exhibit non-constant envelopes with high PAPR. It should also be noted that the method outlined in this paper performs analogue Intermediate Frequency (IF) pre-distortion, as opposed to digital base-band pre-distortion.

OBJECT OF THE INVENTION

It is an object of the invention to overcome or alleviate one or more of the above disadvantages and/or to provide the consumer with a useful or commercial choice.

SUMMARY OF THE INVENTION

In one form, although it need not be the only or indeed the broadest form, the invention resides in a method for linearising a multi-carrier radio frequency transmitter or a multi-user CDMA radio frequency transmitter, including the steps of:

measuring a function of out-of-band signal power in the frequency domain at an output of the radio frequency transmitter; and

applying digital base-band pre-distortion to the radio frequency transmitter according to the measured function of the out-of-band signal power;

wherein the digital base-band pre-distortion is performed by a digital base-band pre-distortion network.

Preferably, digital base-band pre-distortion network coefficients of the digital base-band pre-distortion network are optimised to minimise the measured function of the out-of-band signal power.

Preferably, the digital base-band pre-distortion network coefficients are optimised whilst the transmitter is broadcasting.

Suitably, the digital base-band pre-distortion network is a non-linear behavioural model with memory.

Preferably, the non-linear behavioural model with memory is a pruned Volterra Series.

Suitably, the digital base-band pre-distortion network coefficients are pruned Volterra Series kernel coefficients.

Preferably, the digital base-band pre-distortion network is given by the equation:

$$y[n] = x[n] + \sum_{a=1}^{\left(\frac{P-1}{2}\right)} \left[\sum_{k=0}^{\left\lceil \frac{M+1}{R} \right\rceil - 1} h_{2a+1}[k]\, x[n] \left|x[n]\right|^{2(a-1)} \left|x[n-Rk]\right|^{2} \right]$$

where $h_{2a+1}[k]$ are the digital base-band pre-distortion network kernel coefficients.

Suitably, the memory length M is estimated by:

a) pruning the digital base-band pre-distortion network to a 3rd order single delay digital base-band pre-distortion network given by the equation:

$$y[n] = x[n] + h_3[k]\, x[n]\, |x[n-k]|^2$$

b) Sweeping a delay variable (k) of the 3rd order single delay pre-distortion network from zero upwards; and

c) Observing a value of k when an asymmetry of the transmitter output adjacent channel power spectrum changes wherein the value of k is equal to the memory length M.

Preferably, the function of the out-of-band signal power is a measure of transmitter output non-linearity.

Preferably, the function of the out-of-band signal power involves accumulating a weighted out-of-band power spectral density with respect to frequency.

Preferably, the function of the out-of-band signal power is given by the equation:

$$WACP = \sum_{\mathrm{LAC}\, f} W(f) \times PSD(f) + \sum_{\mathrm{UAC}\, f} W(f) \times PSD(f)$$

Suitably, the weighting function $W(f)$, for either the lower adjacent channel (LAC) or upper adjacent channel (UAC), is a non-increasing function of $|f - f_l|$.

Preferably, the power spectral density is measured with a spectrum analyser.

Preferably, a subset of the digital base-band pre-distortion network kernel coefficients is optimised separately.

Optionally, a combination of 3rd order, a combination of 3rd and 5th order or a combination of 3rd and 5th and 7th order digital base-band pre-distortion network kernel coefficients is optimised separately.

Preferably, the digital base-band pre-distortion network kernel coefficients are optimised according to a local minimum non-gradient based algorithm.

Suitably, the digital base-band pre-distortion network kernel coefficients are optimised according to a global minimum non-gradient based algorithm.

Optionally, the local minimum non-gradient based algorithm is a Nelder-Mead Simplex algorithm.

Preferably, the global minimum non-gradient based algorithm is a Genetic algorithm.

Optionally, a subset of the digital base-band pre-distortion network kernel coefficients, all of the same non-linear order, is optimised separately according to a gradient based algorithm.

Suitably, the gradient based algorithm is a local minimum Gradient Descent algorithm.

BRIEF DESCRIPTION OF THE DRAWINGS

To assist in understanding the invention and to enable a person skilled in the art to put the invention into practical effect, preferred embodiments of the invention will be described by way of example only with reference to the accompanying drawings, in which:

FIG 1 shows a graph of an output spectrum from a prior art radio frequency transmitter;

FIG 2 shows a graph of a frequency response of a prior art output bandpass filter;

FIG 3 shows a block diagram of a radio frequency transmitter according to an embodiment of the present invention.

FIG 4 shows a graph of a weighting function used according to an embodiment of the present invention; and

FIG 5 shows a graph of an output spectrum from a radio frequency transmitter after digital base-band pre-distortion has been applied according to an embodiment of the present invention.

DETAILED DESCRIPTION OF THE INVENTION

Elements of the invention are illustrated in concise outline form in the drawings, showing only those specific details that are necessary to understanding the embodiments of the present invention, but so as not to clutter the disclosure with excessive detail that will be obvious to those of ordinary skill in the art in light of the present description.

In this patent specification, words such as "comprises" or "includes" are not used to define an exclusive set of elements or method steps. Rather, such words merely define a minimum set of elements or method steps included in a particular embodiment of the present invention.

FIG 3 shows a block diagram of a Radio Frequency (RF) transmitter according to an embodiment of the present invention.

Prior art RF transmitters consist of a signal modulator 310 connected directly to a pair of Digital-to-Analogue Converters (DACs) 330, an Inphase-Quadrature (IQ) modulator / frequency upconverter 340, a power amplifier 350, an output bandpass filter 360 and an antenna 365.

The system for linearising the prior art transmitter according to the present invention includes a digital base-band pre-distortion network 320 (connected between the signal modulator 310 and the pair of Digital-to-

Analogue Converters (DACs) 330), a spectral power meter 370, a communication and control module 380 and a mathematical optimiser 390. Also part of the transmitter, but not shown, is an RF directional coupler inserted at the output of the power amplifier 350.

The present invention is designed to linearise the DACs 330, the Inphase-Quadrature (IQ) modulator / frequency upconverter 340 and the power amplifier 350. However a person skilled in the art will realise that the power amplifier 350 is the major source of nonlinearity in an RF transmitter.

The signal modulator 310 generates a multi-carrier OFDM or multi-user CDMA signal (discrete-time, complex (Inphase and Quadrature phase components), base-band) from an incoming bit stream 305. These signals exhibit a non-constant envelope and high Peak to Average Power Ratio. A person skilled in the art will appreciate however that the signal modulator may generate any applicable discrete-time, complex, base-band, communication signal. The output of the signal modulator 310 is input to the digital base-band pre-distortion network 320.

The digital base-band pre-distortion network 320 is a non-linear, discrete-time system operating at base-band and whose non-linear transfer characteristic is designed to be the inverse non-linear transfer characteristic of the combined effects of the DACs 330, the Inphase-Quadrature (IQ) modulator / frequency upconverter 340 and the power amplifier 350. As a result, the cascade of the digital base-band pre-distortion network 320, the DACs 330, the Inphase-Quadrature (IQ) modulator / frequency upconverter 340 and the power amplifier 350 is substantially linear. As stated previously, compared to the DACs 330, the Inphase-Quadrature (IQ) modulator / frequency upconverter 340, the power amplifier 350 is the major source of prior art transmitter nonlinearity.

The non-linear transfer characteristic of the digital base-band pre-distortion network 320 is controlled by adjusting the digital base-band pre-distortion network's coefficients. The digital base-band pre-distortion network

320 is implemented digitally either via a Digital Signal Processor (DSP) or dedicated digital logic. The output of the digital base-band pre-distortion network 320 is a discrete-time, complex, base-band signal.

The pair of Digital-to-Analogue Converters 330 (one for Inphase and the other for Quadrature phase) convert the output of the digital base-band pre-distortion network 320 to continuous-time. The output of the Digital-to-Analogue Converters 330 is thus a continuous-time, complex, base-band signal.

The Inphase-Quadrature (IQ) modulator / frequency upconverter 340 converts the output of the Digital-to-Analogue Converters 330 to a real, Radio Frequency (RF) signal which is input to the power amplifier 350. The power amplifier 350 then amplifies this signal to a broadcast power level. The output from the power amplifier 350 is subsequently filtered by the output bandpass filter 360 to further reduce adjacent channel distortion before being radiated by the antenna 365.

The communication and control module 380 implements the communications link between the digital base-band pre-distortion network 320, spectral-power meter 370 and mathematical optimiser 390. The communications and control module 380 also controls the sequence of events that form the linearisation method. A person skilled in the art will realise that the communications and control module 380 and the mathematical optimiser 390 are implemented together in software with suitable hardware.

The spectral power meter 370 is connected to the output of the power amplifier 350 via an RF directional coupler (not shown). The spectral power meter 370 may be a spectrum analyser. However it should be appreciated that a spectral power measurement may be made using a dedicated circuit or any other suitable device. The spectral power meter 370 measures Power Spectral Density (PSD) at the output of the power amplifier 350 at a frequency specified by the mathematical optimiser 390 (and communicated via the communication and control module 380).

A person skilled in the art will realise that the spectral power meter 370 may also be connected at the output from the output bandpass filter 360 or at an output of any other component that may be connected between the power amplifier 350 and the antenna 365 which are also considered in this specification to be the output of the transmitter.

From multiple PSD measurements taken by the spectral power meter 370 at various out-of-band frequencies, the mathematical optimiser 390 computes a function of the out-of-band signal power. The function of the out-of-band signal power represents a frequency domain measure of the transmitter output non-linearity. The mathematical optimiser 390 then optimises the coefficients of the digital base-band pre-distortion network 320 (via the communication and control module 380) according to the function of the out-of-band signal power in order to linearise the transmitter.

In one embodiment, the function of the out-of-band signal power is an Adjacent Channel Power (ACP) measurement. However it should be appreciated that other functions of the out-of-band signal power may be used. The ACP is computed by accumulating PSD measurements made at different out-of-band frequencies using the spectrum analyser. In one embodiment, the resolution bandwidth over which the PSD is measured and the video averaging that is applied to the PSD measurement, is varied depending on the type of modulation output from the signal modulator 310 and a type of spectrum analyser used, as would be understood by a person skilled in the art.

ACP is given by the equation:

$$ACP = \int_{\mathrm{LAC}\,f} PSD(f)df + \int_{\mathrm{UAC}\,f} PSD(f)df \qquad \text{Eq.1}$$

where:

$PSD(f)$ is the transmitter output power spectral density as a function of frequency;

LAC is one or more lower adjacent channels; and

UAC is one or more upper adjacent channels.

ACP is considered a pure measure of transmitter output non-linearity. The ACP measure of transmitter output non-linearity assumes that distortion produced at each out-of-band frequency in the transmission is equally detrimental. However, distortion at some frequencies may be considered more detrimental than others. For instance, distortion at out-of-band frequencies close to the band edges of the allocated transmission channel may be considered the most detrimental because the output bandpass filter 360 has less attenuation there as shown in FIG 2.

In order to place more emphasis on reducing the distortion at out-of-band frequencies close to the allocated transmission band edges, the ACP measure of transmitter output non-linearity may be refined with a frequency dependent weighting to give the Weighted Adjacent Channel Power (WACP):

$$WACP = \int_{\mathrm{LAC}\, f} W(f) \times PSD(f)\,df + \int_{\mathrm{UAC}\, f} W(f) \times PSD(f)\,df \qquad \text{Eq.2}$$

where:

$W(f)$ is a non-negative, frequency dependent weighting function;

$PSD(f)$ is the transmitter output power spectral density as a function of frequency;

LAC is one or more lower adjacent channels; and

UAC is one or more upper adjacent channels.

ACP is the specific case of WACP when the weighting function $W(f) = 1$. Compared to ACP, WACP is a more general measure of transmitter output non-linearity. WACP is considered non analytic as it is derived from spectrum analyser PSD measurements rather than formularised.

In practice it is not possible to integrate. Rather small discrete steps in frequency are summed resulting in Eq.3 below:

$$WACP = \sum_{\mathrm{LAC}\,f} W(f) \times PSD(f) + \sum_{\mathrm{UAC}\,f} W(f) \times PSD(f) \qquad \text{Eq.3}$$

It should be noted that only the out-of-band frequencies for which the PSD is above the spectrum analyser noise floor need to be included in the summation in Eq. 3.

Multi-carrier OFDM and multi-user CDMA signals are considered random processes due to the random nature of the input bit stream 305. As a result, the transmitter output signal is also considered a random process and the WACP measure must be modelled as a random variable with a mean and a spread. It should be appreciated that taking several WACP samples and averaging may give a better estimate compared to taking a single WACP sample alone. However by choosing robust optimisation algorithms (discussed later), the detrimental effects of WACP randomness can be mitigated and the amount of averaging reduced.

The weighting functions of Eq.3 that are of particular practical importance are those which place greater weighting at out-of-band frequencies closer to the allocated transmission band edges where the attenuation of the output bandpass filter 360 is reduced. The weighting functions are non-increasing functions of $|f - f_l|$ where for the lower adjacent channel frequencies f_l is the lower edge frequency of the allocated transmission band and for the upper adjacent channel frequencies f_l is the upper edge frequency of the allocated transmission band. An example of such a weighting function is shown graphically in FIG 3 however it should be appreciated that there are many such weighting functions and some examples are given in equations 4 to 8 below:

$$W(f) = C \qquad \text{Eq.4}$$

$$W(f) = \left(\frac{C}{|f - f_I|}\right) \qquad \text{Eq.5}$$

$$W(f) = \left(\frac{C}{|f - f_I|}\right)^2 \qquad \text{Eq.6}$$

$$W(f) = \begin{cases} 0 & \text{for } |f - f_I| > |f_O - f_I| \\ \left(\frac{-W_I|f - f_I|}{|f_O - f_I|}\right) + W_I & \text{for } |f - f_I| \le |f_O - f_I| \end{cases} \qquad \text{Eq.7}$$

$$W(f) = \begin{cases} 0 & \text{for } |f - f_I| > |f_O - f_I| \\ \left(\frac{W_I|f - f_I|^2}{|f_O - f_I|^2}\right) - \left(\frac{2W_I|f - f_I|}{|f_O - f_I|}\right) + W_I & \text{for } |f - f_I| \le |f_O - f_I| \end{cases}$$

Eq.8

where:

C is a positive constant;

f_I is the transmitter's allocated transmission band 470 edge frequency (a lower edge 410 for the lower adjacent channel 420 and an upper edge 430 for the upper adjacent channel 440);

f_O is an outer frequency 450 (further from the carrier than f_I) at which the weighting function falls to zero; and

W_I is a desired weighting 460 at f_I.

The digital base-band pre-distortion network 320 is based on a suitable non-linear base-band transmitter model. A behavioural model rather than a circuit level model is chosen in order to ensure the pre-distortion network 320 is more generally applicable. Given that the power amplifier 350 with a wideband input signal (DVB-T$\approx$7 MHz, DAB$\approx$1.5 MHz, WCDMA$\approx$5

MHz) exhibits substantial non-linear memory, the behavioural model must also possess memory. A transmitter is said to have memory if its output is a function of the past inputs. Transmitter memory manifests itself as asymmetry between the lower and upper adjacent channel power spectral densities.

Some non-linear behavioural models with memory include Neural Networks, Hammerstein/Weiner filters and the Volterra Series. Note narrowband memory-less AM-AM/AM-PM models are not suitable as the transmission modulation bandwidth is wideband in nature. In one embodiment, the Volterra Series model is chosen as it is the most general. However it should be appreciated that other models may be used.

A discrete-time, causal, complex base-band Volterra Series with maximum non-linearity P (odd) and memory M representing the digital base-band pre-distortion network 320 is given by the equation:

$$y[n]=\sum_{k_1=0}^{M}h_1[k_1]x[n-k_1]+\sum_{a=1}^{\left(\frac{P-1}{2}\right)}\left[\sum_{k_1=0}^{M}\cdots\sum_{k_{2a+1}=0}^{M}\left(h_{2a+1}[k_1,\cdots,k_{2a+1}]\prod_{i=1}^{a+1}x[n-k_i]\prod_{j=a+2}^{2a+1}x^*[n-k_j]\right)\right]$$

Eq.9

where:

$x[n]$ is the input signal complex envelope (the signal output from the signal modulator 310);

$y[n]$ is the output signal complex envelope (the signal input to the Digital-to-Analogue Converters 330);

M is memory;

P is the maximum order of non-linearity (odd);

k is a delay variable;

$x^*[\cdot]$ denotes complex conjugation; and

$h_{2a+1}[k_1,\cdots,k_{2a+1}]$ is called the $(2a+1)^{th}$ order Volterra kernel (or pre-distortion network kernel) and the entire set of kernels $a=1$ to $(P-1)/2$ fully characterises the pre-distortion network. It should be noted that the above Volterra Series only contains odd ordered kernels due to the channel selectivity of the output bandpass filter 360. It is also noted that the kernels are complex containing real and imaginary parts.

In one embodiment, the digital base-band pre-distortion network 320 of the present invention is based on the Volterra Series given in Eq.9. However the number of coefficients of the pre-distortion network kernel to be estimated by optimisation can be too large. The number of coefficients to be estimated (or overall kernel size of the Volterra Series) increases exponentially with the degree of non-linearity P and memory length M. As a result, it is desirable to prune the Volterra Series of Eq. 9 in order to reduce the size of the pre-distortion network kernel.

There are many pruning techniques such as Memory Polynomial, Dynamic Deviation Reduction, Physical Knowledge, Near Diagonality Restriction, Base-Band derived Volterra Model and Volterra Behavioral Wideband pruning that may used. Of these the Volterra Behavioral Wideband pruning technique is used given it is based on wideband signal theory and offers a good trade-off between kernel size and performance. However it should be appreciated that other pruning techniques may be used. Applying the Volterra Behavioral Wideband pruning technique to Eq.9 results in the following pruned Volterra Series:

$$y[n]=h_1x[n]+\sum_{a=1}^{\left(\frac{P-1}{2}\right)}\left[\sum_{k_1=0}^{M}\cdots\sum_{k_a=0}^{M}\left(h_{2a+1}[k_1,\cdots,k_a]x[n]\prod_{j=1}^{a}\left|x[n-k_j]\right|^2\right)\right] \qquad \text{Eq.10}$$

Although pruning Eq.10 substantially reduces the size of the pre-distortion network kernel compared to Eq.9, the number of coefficients to be estimated via optimisation is still large, particularly for higher orders. As a result, while it should be appreciated that Eq.10 may be used, it is desirable to apply further pruning by restraining dynamics (the number of delayed input terms $x[n-k]$) to 2^{nd} order. Also, given the pre-distortion network 320 is not required to perform linear compensation, the linear pre-distortion kernel h_1 may be removed. These additional pruning steps lead to:

$$y[n] = x[n] + \sum_{a=1}^{\left(\frac{P-1}{2}\right)} \left[\sum_{k=0}^{M} h_{2a+1}[k]\, x[n]\, |x[n]|^{2(a-1)}\, |x[n-k]|^{2} \right] \qquad \text{Eq.11}$$

Although it should be appreciated that Eq.11 may be used, a final stage of pruning can be performed as the input signal to the digital base-band pre-distortion network 320 is heavily oversampled. Oversampling by at least the highest pre-distortion network non-linearity should occur in order to account for spectral regrowth added by the pre-distortion network 320 and therefore avoid discrete-time spectral aliasing. The oversampling leads to an input signal with a very narrow discrete-time spectral bandwidth given by BW / f_s where BW represents the input signal continuous-time spectral bandwidth and f_s represents the sampling rate. As a result, the change between adjacent input signal samples can be considered very small to the point where groups of R input samples (R being small) can be assumed equal. With this assumption, the pre-distortion network pruned Volterra Series of Eq.11 can be refined with R-sample delay increments instead of single-sample delay increments without loss in performance as follows:

$$y[n] = x[n] + \sum_{a=1}^{\left(\frac{P-1}{2}\right)} \left[\sum_{k=0}^{\left\lceil \frac{M+1}{R} \right\rceil - 1} h_{2a+1}[k]\, x[n]\, |x[n]|^{2(a-1)}\, |x[n-Rk]|^{2} \right] \qquad \text{Eq.12}$$

Eq.12 represents the final digital base-band pre-distortion network 320 derived from the pruned Volterra Series. Although the pre-distortion network 320 has been refined to operate with internal R-sample delay increments, the pre-distortion network 320 is clocked at the oversampling rate to avoid spectral regrowth aliasing at the output of the pre-distortion network 320. With these larger R-sample delay increments, afforded by the input signal's very narrow discrete-time spectral bandwidth, the pre-distortion network 320 has a greater computational efficiency and the pre-distortion network kernel is further pruned by an extra approximate factor of R. It should be appreciated that for the case R = 1, the digital base-band pre-distortion network of Eq.12 reduces to that of Eq.11. The value of R may be estimated from the input signal's discrete-time spectral bandwidth. The smaller the discrete-time spectral bandwidth, the greater R may be. Example discrete-time spectral bandwidths and corresponding values of R are shown in the table below:

Discrete-Time Bandwidth (BW / f_s)	R
0.022888	4
0.054253	3
0.109375	2

It should be appreciated that a conservative estimate of R is made. Over estimating R in order to provide extra kernel pruning may result in an invalid assumption that "groups of R input samples are equal" and hence degraded pre-distortion network performance.

Memory length M in Eq.12 of the pre-distortion network 320 is determined experimentally as follows. The pre-distortion network 320 is pruned to a 3rd order single delay pre-distortion network as shown in Eq.13 below:

$$y[n] = x[n] + h_3[k]\, x[n]\, |x[n-k]|^2 \qquad \text{Eq.13}$$

The delay k of Eq.13 is swept from zero upwards. Whilst performing the sweep, $h_3[k]$ is chosen such that there is a small but observable change in the level of the measured output adjacent channel power spectrum. The asymmetry of the transmitter output adjacent channel power spectrum is observed for changes prior to applying the pre-distortion, and the value of delay k corresponding to the change in asymmetry is chosen as the memory length M. While this experimental approach for estimating M performs well, it should be appreciated that other schemes for estimating M may be used instead.

Maximum non-linearity P in Eq.12 of the pre-distortion network 320 is set to 9. This is a result of the transmitter's dominant 3rd order non-linearity and hence the significant 5th, 7th and 9th order parasitic non-linearities generated from the 3rd order pre-distortion process. That is, the 5th, 7th and 9th order pre-distortion network kernels are predominantly used to compensate for the 5th, 7th and 9th order distortion introduced by the 3rd order pre-distortion network kernel. However it should be appreciated that larger or smaller values of P can be used depending on the transmitter and performance requirements.

In the final digital base-band pre-distortion network of Eq. 12, $h_{2a+1}[k]$ is called the $(2a+1)^{\text{th}}$ order pre-distortion network kernel and the entire set of kernels $h_3[k]$, $h_5[k]$, $h_7[k]$ and $h_9[k]$ fully characterises the pre-distortion network 320. This set of kernels, with expanded k, represents complex coefficients of the pre-distortion network which are to be optimised. The pre-distortion network coefficients can thus be represented mathematically as a vector:

$$\underline{h} = \{\underline{h}_3 \quad \underline{h}_5 \quad \underline{h}_7 \quad \underline{h}_9\} \text{ with } \underline{h}_z = h_z[k] \text{ for } k = 0 \text{ to } \left\lceil \tfrac{M+1}{R} \right\rceil - 1$$

Computation of the pre-distortion network kernel may now be modelled as a single objective mathematical optimization problem with:

- A Vector Space to Be Optimized:

$$\underline{h} = \{\underline{h}_3 \quad \underline{h}_5 \quad \underline{h}_7 \quad \underline{h}_9\} \text{ with } \underline{h}_z = h_z[k] \text{ for } k = 0 \text{ to } \left\lceil \tfrac{M+1}{R} \right\rceil - 1$$

- An Objective Function to be minimised (derived from the PSD measurement taken from the spectrum analyser):

$$WACP = \sum_{\mathrm{LAC}\,f} W(f) \times PSD(f) + \sum_{\mathrm{UAC}\,f} W(f) \times PSD(f)$$

The pre-distortion network kernel $\underline{h}_{\mathrm{OPTIMAL}}$ minimizes the Weighted Adjacent Channel Power in order to linearise the transmitter.

Optimisation of the vector space $\underline{h}$ is performed by the mathematical optimiser 390. It has been found that a single mathematical optimisation over the entire vector space $\underline{h}$ leads to below average likelihood of convergence given the poor scaling (or large difference in magnitude) that exists between kernel coefficients of different non-linear orders. Although it should be appreciated that performing a single optimisation over the entire vector space $\underline{h}$ may be performed, it is preferable to optimise the vector space $\underline{h}$ over several separate optimisations, each optimisation focused on a subset $\underline{h}_{SUB}$ of the vector space $\underline{h}$. That is $\underline{h}_{SUB} \subseteq \underline{h}$.

In some embodiments the mathematical optimiser 390 used to optimise $\underline{h}_{SUB}$ may be either Gradient based or non-Gradient based (for example a Direct Search or Stochastic optimiser).

Gradient based optimisers require knowledge of the WACP objective function 1st order derivative characteristics (Gradient vector) and possibly 2nd order derivative characteristics (Hessian matrix). In practice, the 1st and 2nd order derivative characteristics are approximated using Finite Differences (Gradient or Hessian) or a Symmetric-Rank-1 update (Hessian). Whilst Gradient based optimisation is technically superior to other forms of optimisation, it is known to be computationally intensive and susceptible to measurement noise.

The following Gradient based mathematical optimisers have been tested for their suitability:

- Gradient Descent (a local mathematical optimiser requiring Gradient vector computation);
- Trust Region Newton (a local mathematical optimiser requiring both

Gradient vector and Hessian matrix computation); and

- Alpha Branch & Bound (a global mathematical optimiser requiring both Gradient vector and Hessian matrix computation).

The following observations were made as a result of the tests:

1. The random nature of the WACP objective function leads to trust region uncertainty and in certain cases Gradient / Hessian approximation error;
2. Poor scaling (or a large difference in magnitude) existing between kernel coefficients of different non-linear order leads to extreme optimisation step size sensitivity; and
3. Computation of the Hessian matrix was impractically slow when considering the large number of matrices required to be computed over the entire optimisation.

Observation 3 above suggests that the Trust Region Newton and Alpha Branch & Bound optimisers (both requiring Hessian matrix computation) are not preferred. The Gradient Descent optimiser is recommended, but as indicated by observation 2 above, only when all elements of $\underline{h}_{SUB}$ are kernel coefficients of the same non-linear order for example the 3rd order, where coefficients are well scaled and optimiser step size is insensitive. Observation 1 above suggests that when performing any Gradient based optimisation, WACP averaging is recommended in order to reduce Gradient vector approximation error caused by the random nature of the WACP objective function.

In contrast to Gradient based mathematical optimisers, Non-Gradient based mathematical optimisers rely solely on knowledge of the WACP objective function value. That is, 1st and 2nd order derivative characteristics are not required. The WACP objective function value is measured directly. Whilst non-Gradient based optimisation is not as technically apt as Gradient based optimisation, it is less computationally intensive and is less susceptible to measurement noise, making it overall more robust. Direct Search and Stochastic algorithms are particular examples of non-Gradient based mathematical optimisation algorithms.

The following non-Gradient based mathematical optimisers have been tested for their suitability:

- Nelder-Mead Simplex (a Local, Direct Search mathematical optimiser); and
- Genetic (a Global, Stochastic mathematical optimiser);

The following observations were made as a result of the tests:

1. The random nature of the WACP objective function did not show any signs of degrading optimisation performance;
2. Poor scaling (or a large difference in magnitude) existing between kernel coefficients of different non-linear order may be accommodated by individual non-linear order increments / variance; and
3. Computation was highly efficient.

It can be seen from these observations that the Nelder-Mead Simplex optimiser is suitable for use as the local mathematical optimiser and the Genetic optimiser is suitable for use as the global mathematical optimiser. However a person skilled in the art will appreciate that other mathematical optimisers, for example Simulated Annealing, may be used. Furthermore WACP averaging can be avoided when using non-Gradient based optimisation. This is due to the robustness of the optimisers.

When using the Nelder-Mead Simplex optimiser, it may be necessary to restart the optimiser periodically in order to reset its simplex (an N+1 point constellation on the objective function surface, where N is the number of elements to optimise) and avoid convergence at a poor local minima.

Also the Genetic optimiser's current progress in locating the global minima can be monitored by comparing chromosomes from the fittest population. For each chromosome of the fittest population, Genes are laid across the x-axis and the corresponding Gene values are plotted on the y-axis. If chromosomes show varying Gene values, the optimiser is still in the process of locating the global minima and should be left to continue. Alternatively, if all chromosomes show similar Gene values, the optimiser has honed in onto the global minima and the optimisation can be ceased. At this point it is then recommended to refine the output of the Genetic optimiser by applying a follow up Nelder-Mead Simplex local optimisation.

There are two phases for optimising the kernel coefficients of the digital base-band pre-distortion network 320, being the Initial Optimisation phase (when the transmitter has been initially installed) and the Adaptive Optimisation phase (when the transmitter is operational).

The Initial Optimisation phase involves computing initial coefficients of the pre-distortion network kernel when the transmitter is first commissioned. The initial coefficients are computed with the output of the output bandpass filter 360 connected to a dummy load rather than being broadcast via the antenna 365. This is because out-of-band signal power will exceed a regulatory spectral mask until the coefficients of the pre-distortion network kernel have been initially optimised. Once the Initial Optimisation phase has been completed and a regulatory spectral mask has been met, the transmitter is ready for broadcasting and the output of the output bandpass filter 360 can be connected to the antenna 365.

Over time, the transmitter's non-linear transfer characteristics will drift slowly due to component aging (transistors and capacitors), temperature fluctuations and power supply voltage variations. Thus the coefficients of the pre-distortion network kernel computed during the Initial Optimisation phase do not remain optimal over the entire lifetime of the transmitter. Hence the need for the Adaptive Optimisation phase.

The Adaptive Optimisation phase adapts the coefficients of the pre-distortion network kernel, in order to maintain optimality when the transmitter's non-linear transfer characteristics change. The Adaptive Optimisation phase occurs whilst the transmitter is broadcasting a normal signal via the antenna, as taking the transmitter off-air is undesirable. Injecting known test signals into the transmitter is not necessary. All adaption is based on the transmitter's normal signal.

The Initial Optimisation and Adaptive Optimisation phases are described below. As discussed previously, in order to improve optimiser likelihood of convergence, it is preferable to optimise the vector space $\underline{h}$ over several separate optimisations, each optimisation focused on a subset $\underline{h}_{SUB}$ of the vector space $\underline{h}$. At all times throughout the Initial Optimisation and Adaptive Optimisation phases, $\underline{h}_{SUB}$ is chosen to be that subset of $\underline{h}$ which has an immediately dominant effect on reducing the WACP objective function. For example, if the WACP objective function is comprised of dominant x^{th} and y^{th} order non-linear distortion power, then $\underline{h}_{SUB}$ is chosen to be $\underline{h}_{SUB} = \{\underline{h}_x \quad \underline{h}_y\}$ where $\underline{h}_x$ and $\underline{h}_y$ are the x^{th} and y^{th} order pre-distortion network kernel coefficients respectively. In addition, in cases where poor scaling (a large difference in magnitude) exists between coefficients of $\underline{h}_{SUB}$, $\underline{h}_{SUB}$ is split into separate subsets each with improved coefficient scaling and separate optimisations are performed on these separate subsets.

In one embodiment the Initial Optimisation phase is performed according to the following schedule. However a person skilled in the art will realise that there are many permutations and combinations of initially optimising the coefficients of the pre-distortion network kernel.

1st Optimisation:

Order of pre-distortion network kernel to be optimised	**Measurement to be minimised**	**Weighting function**	**Optimiser used**
3rd order real part	WACP	non-increasing function of $\lvert f - f_I \rvert$	Global
3rd order imaginary part	WACP	non-increasing function of $\lvert f - f_I \rvert$	Global

2nd Optimisation:

Order of pre-distortion network kernel to be optimised	**Measurement to be minimised**	**Weighting function**	**Optimiser used**
5th, 7th, 9th order real part	WACP	non-increasing function of $\lvert f - f_I \rvert$	Global
5th, 7th, 9th order imaginary part	WACP	non-increasing function of $\lvert f - f_I \rvert$	Global

3rd Optimisation:

Order of pre-distortion network kernel to be optimised	**Measurement to be minimised**	**Weighting function**	**Optimiser used**
3rd order real part	WACP	non-increasing function of $\lvert f - f_l \rvert$	Local
3rd order imaginary part	WACP	non-increasing function of $\lvert f - f_l \rvert$	Local

4th Optimisation:

Order of pre-distortion network kernel to be optimised	**Measurement to be minimised**	**Weighting function**	**Optimiser used**
5th, 7th, 9th order real part	WACP	non-increasing function of $\lvert f - f_l \rvert$	Local
5th, 7th, 9th order imaginary part	WACP	non-increasing function of $\lvert f - f_l \rvert$	Local

In one embodiment the Adaptive Optimisation phase is performed according to the following schedule. Again, a person skilled in the art will realise that there are many permutations and combinations of adaptively optimising the coefficients of the pre-distortion network kernel.

1st Optimisation:

Order of pre-distortion network kernel to be optimised	**Measurement to be minimised**	**Weighting function**	**Optimiser used**
3rd order real part	WACP	non-increasing function of $\|f-f_I\|$	Local
3rd order imaginary part	WACP	non-increasing function of $\|f-f_I\|$	Local

2nd Optimisation:

Order of pre-distortion network kernel to be optimised	**Measurement to be minimised**	**Weighting function**	**Optimiser used**
5th, 7th, 9th order real part	WACP	non-increasing function of $\|f-f_I\|$	Local
5th, 7th, 9th order imaginary part	WACP	non-increasing function of $\|f-f_I\|$	Local

The Adaptive Optimisation schedule is repeated indefinitely, or when the WACP is observed to increase, in order to maintain coefficient optimality and ensure the out-of-band signal power remains within the spectral mask. In some embodiments, the 5th, 7th and 9th order coefficients of the pre-distortion network kernel are optimised at the same time or in parallel, however it should be appreciated that the 5th, 7th and 9th order coefficients may be optimised separately or sequentially.

In one embodiment, a combination of 3rd order, a combination of 3rd and 5th order or a combination of 3rd and 5th and 7th order pre-distortion network kernel coefficients are optimised separately.

FIG 5 shows a graph 500 of an output spectrum from the transmitter before the application of the digital base-band pre-distortion network 320 and after the digital base-band pre-distortion network 320 has been applied and optimised. Trace 502 (circular markers) shows the output from the transmitter before the application of the pre-distortion network 320 and trace 504 (triangular markers) shows the output from the transmitter when the pre-distortion network 320 has been applied and optimised. As can be seen in FIG5, co-channel distortion 506 and adjacent channel distortion 508 may be reduced.

The two main approaches to pre-distortion network kernel computation are *Direct/Indirect Learning* and *Model Based Inversion*.

The Direct/Indirect Learning approach treats pre-distortion network kernel computation as a parameter estimation problem; specifically a linear regression problem solved using Least Mean Squares (LMS) adaption. The Direct/Indirect Learning approach exhibits the following problems:

- The LMS error criterion on which to adapt is obtained via a time domain feedback path (from output to input). This feedback path must compensate for amplifier gain and propagation time delay (both frequency dependent) as well as perform analogue-to-digital conversion. In practice, gain/delay compensation error and frequency dependent Analogue-to-Digital Converter distortion is present, ultimately leading to suboptimal performance.

- Performing the LMS adaption is computationally intensive (many digital multiplications). This is a result of the linear redefinition of the non-linear pre-distortion network (data input) during linear regression modelling.
- In the specific case of Direct Learning, the error criterion surface is assumed quadratic. However, this assumption is incorrect given the transmitter's non-linearity. As a result, global convergence of the error criterion is not guaranteed using local LMS optimisation.
- In the specific case of Indirect Learning, post-distortion parameters are first estimated and then translated to pre-distortion parameters. Applying this translation in non-linear systems is not properly formal in the mathematical sense and leads to approximation error.

The method of the present invention may be classified as a parameter estimation technique but differs from the Direct/Indirect Learning approach in the following ways:

Technique of the Present Invention	Direct/Indirect Learning Technique
Pre-distortion network kernel computation is modelled as a generic optimisation problem.	Pre-distortion network kernel computation is modeled as a specific linear regression problem solved using LMS adaption.
The objective function to be minimised is a pure, frequency domain WACP.	The error criterion to be minimised is a time domain feedback signal exhibiting frequency dependent gain/delay compensation error and ADC distortion.

Optimisation is performed via non-Gradient based algorithms which have minimal computational intensity.	Optimisation is performed via the LMS algorithm which becomes computationally intensive with the linear redefinition of the non-linear pre-distortion network (linear regression modelling).
The objective function is assumed to have many local minima. As a result, both global and local optimisation algorithms are appropriately employed to find the global minimum.	Direct Learning uses the local LMS optimiser on an incorrectly assumed quadratic error surface. The optimiser may thus converge on a local minimum rather than the global minimum and thus result in a suboptimal performance.
Post-distortion is avoided and therefore so too are translation errors.	Indirect Learning involves post-distortion and translation. Translation leads to approximation error.

The second main approach to pre-distortion kernel computation is Model Based Inversion. As its name suggests, this approach involves choosing a blank behavioural model for the non-linear transmitter, deriving model parameters via direct measurement (system identification) and then mathematically inverting the model to obtain the pre-distortion network. It logically follows that for this approach, the maximum linearisation performance is limited by the accuracy of the transmitter model and the accuracy of the inversion. While this approach has proven successful for narrowband modulating signals with an AM-AM/AM-PM transmitter model, it is not well suited to the wideband case. This is because as signal bandwidth increases, it becomes increasingly difficult to accurately model the transmitter's frequency dependent characteristics and memory effects. As a result, for the wideband signals used in modern transmission formats, larger inaccuracies exist in the transmitter model and hence the mathematically inverted pre-distortion network. This ultimately leads to poor linearisation performance.

Another problem exists with this Model Based Inversion approach. In general, the system identification techniques used to derive transmitter model parameters require known test signals (possessing desirable characteristics) to be injected into the transmitter. In practice, this means that any time the pre-distortion network coefficients are to be updated (in order to track changes in the transmitter's non-linear transfer characteristics occurring over time), the transmitter must be taken off-air so that the known test signals can be injected in place of the normal broadcast signal. Given an off-air transmitter is highly undesirable for the transmitter operators, the pre-distortion network coefficients are rarely updated, despite needing to be, and hence the transmitter is poorly linearised for the majority of its operational life.

Given the present invention does not require transmitter modelling or inversion, it does not suffer from the problems inherent with the Model Based Inversion approach.

Thus the method and system of the present invention for linearising a radio frequency transmitter has many advantages over the prior art including:

1) The pre-distortion method is modelled as a generic single objective mathematical optimisation problem. As a result, all techniques of the well established field of mathematical optimisation can be drawn upon to find the best solution, both globally and locally. This is opposed to modelling the problem as a specific linear regression problem, incorrectly assuming a single local minimum and relying on the LMS algorithm.

2) The pre-distortion method performs adaptive optimisation based on a frequency domain measure of transmitter output non-linearity

which does not require a full feedback path and is hence error free. This is in direct contrast to a time domain measure which requires a full feedback path and hence exhibits feedback gain/delay compensation error and ADC distortion.

3) The pre-distortion method avoids transmitter modelling and inversion and hence the associated signal bandwidth limitations.

4) The digital base-band pre-distortion network is a pruned Volterra Series with memory:

   a. Possessing memory means that the pre-distortion network is well suited to the wideband signals (multi-carrier OFDM and multi-user CDMA) used in modern communication systems.

   b. Pruning reduces the kernel size of the pre-distortion network and therefore makes it well suited to mathematical optimisation.

5) The pre-distortion method possesses a simple, repeatable optimisation schedule for both the Initial Optimisation and Adaptive Optimisation phases.

6) The digital base-band pre-distortion network is able to adapt to changes in the transmitter's non-linear transfer characteristics (occurring over time) without having to take the transmitter off-air (Adaptive Optimisation phase). As a result, the transmitter is both on-air and optimally linearised for its entire operational life.

7) The pre-distortion method uses robust non-Gradient based optimisation algorithms and therefore requires minimal computational processing.

8) The pre-distortion method uses both global and local optimisation algorithms where appropriate and thus has a high likelihood of convergence to the correctly assumed global minimum.

9) The only measurement hardware required is a standard spectrum

analyser (or spectral power meter). No signal phase measurement is necessary.

10) The pre-distortion method may be applied to digital television (DVB-T), digital radio (DAB), 3rd Generation mobile (WCDMA) and 4th Generation mobile (OFDMA) signal formats, all wideband with non-constant envelope and high PAPR.

11) The pre-distortion method works at different carrier frequencies thereby making it suitable for the entire radio frequency transmission band.

12) Apart from standard spectrum analyser (or spectral power meter) calibration, no additional calibration/maintenance is required given a full time domain feedback path is avoided.

13) The process is fully automated and therefore field technician friendly.

The above description of various embodiments of the present invention is provided for purposes of description to one of ordinary skill in the related art. It is not intended to be exhaustive or to limit the invention to a single disclosed embodiment. As mentioned above, numerous alternatives and variations to the present invention will be apparent to those skilled in the art of the above teaching. Accordingly, while some alternative embodiments have been discussed specifically, other embodiments will be apparent or relatively easily developed by those of ordinary skill in the art. Accordingly, this patent specification is intended to embrace all alternatives, modifications and variations of the present invention that have been discussed herein, and other embodiments that fall within the spirit and scope of the above described invention.

Limitations in any patent claims associated with the present disclosure should be interpreted broadly based on the language used in the claims, and such limitations should not be limited to specific examples described herein. In this specification, the terminology "present invention" is used as a reference to one or more aspects within the present disclosure. The terminology "present invention" should not be improperly interpreted as an identification of critical elements, should not be improperly interpreted as applying to all aspects and embodiments, and should not be improperly interpreted as limiting the scope of any patent claims.

CLAIMS

1. A method for linearising a multi-carrier radio frequency transmitter or a multi-user CDMA radio frequency transmitter, including the steps of:

measuring a function of out-of-band signal power in the frequency domain at an output of the radio frequency transmitter; and

applying digital base-band pre-distortion to the radio frequency transmitter according to the measured function of the out-of-band signal power;

wherein the digital base-band pre-distortion is performed by a digital base-band pre-distortion network.

2. The method of claim 1 wherein digital base-band pre-distortion network coefficients of the digital base-band pre-distortion network are optimised to minimise the measured function of the out-of-band signal power.

3. The method of claim 2 wherein the digital base-band pre-distortion network coefficients are optimised whilst the transmitter is broadcasting.

4. The method of claim 1 wherein the digital base-band pre-distortion network is a non-linear behavioural model with memory.

5. The method of claim 4 wherein the non-linear behavioural model with memory is a pruned Volterra Series.

6. The method of claim 2 wherein the digital base-band pre-distortion network coefficients are pruned Volterra Series kernel coefficients.

7. The method of claim 1 wherein the digital base-band pre-distortion network is given by the equation:

$$y[n] = x[n] + \sum_{a=1}^{\left(\frac{P-1}{2}\right)} \left[\sum_{k=0}^{\left\lceil \frac{M+1}{R} \right\rceil - 1} h_{2a+1}[k]\, x[n] \left| x[n] \right|^{2(a-1)} \left| x[n-Rk] \right|^2 \right]$$

where $h_{2a+1}[k]$ are the digital base-band pre-distortion network kernel coefficients.

8. The method of claim 7 wherein the memory length M is estimated by:

a) pruning the digital base-band pre-distortion network to a 3rd order single delay digital base-band pre-distortion network given by the equation:

$$y[n] = x[n] + h_3[k]\, x[n] \left| x[n-k] \right|^2$$

b) Sweeping a delay variable (k) of the 3rd order single delay pre-distortion network from zero upwards; and

c) Observing a value of k when an asymmetry of the transmitter output adjacent channel power spectrum changes wherein the value of k is equal to the memory length M.

9. The method of claim 1 wherein the function of the out-of-band signal power is a measure of transmitter output non-linearity.

10. The method of claim 9 wherein the function of the out-of-band signal power involves accumulating a weighted out-of-band power spectral density with respect to frequency.

11. The method of claim 10 wherein the function of the out-of-band signal power is given by the equation:

$$WACP = \sum_{\text{LAC}\, f} W(f) \times PSD(f) + \sum_{\text{UAC}\, f} W(f) \times PSD(f)$$

12. The method of claim 11 wherein the weighting function $W(f)$, for either the lower adjacent channel (LAC) or upper adjacent channel (UAC), is a non-increasing function of $|f - f_l|$.

13. The method of claim 10 or claim 11 wherein the power spectral density is measured with a spectrum analyser.

14. The method of claim 7 wherein a subset of the digital base-band pre-distortion network kernel coefficients is optimised separately.

15. The method of claim 14 wherein a combination of 3rd order, a combination of 3rd and 5th order or a combination of 3rd and 5th and 7th order digital base-band pre-distortion network kernel coefficients is optimised separately.

16. The method of claim 14 wherein the digital base-band pre-distortion network kernel coefficients are optimised according to a local minimum non-gradient based algorithm.

17. The method of claim 14 wherein the digital base-band pre-distortion network kernel coefficients are optimised according to a global minimum non-gradient based algorithm.

18. The method of claim 16 wherein the local minimum non-gradient based algorithm is a Nelder-Mead Simplex algorithm.

19. The method of claim 17 wherein the global minimum non-gradient based algorithm is a Genetic algorithm.

20. The method of claim 14 wherein a subset of the digital base-band pre-distortion network kernel coefficients, all of the same non-linear order, is optimised separately according to a gradient based algorithm.

21. The method of claim 20 wherein the gradient based algorithm is a local minimum Gradient Descent algorithm.

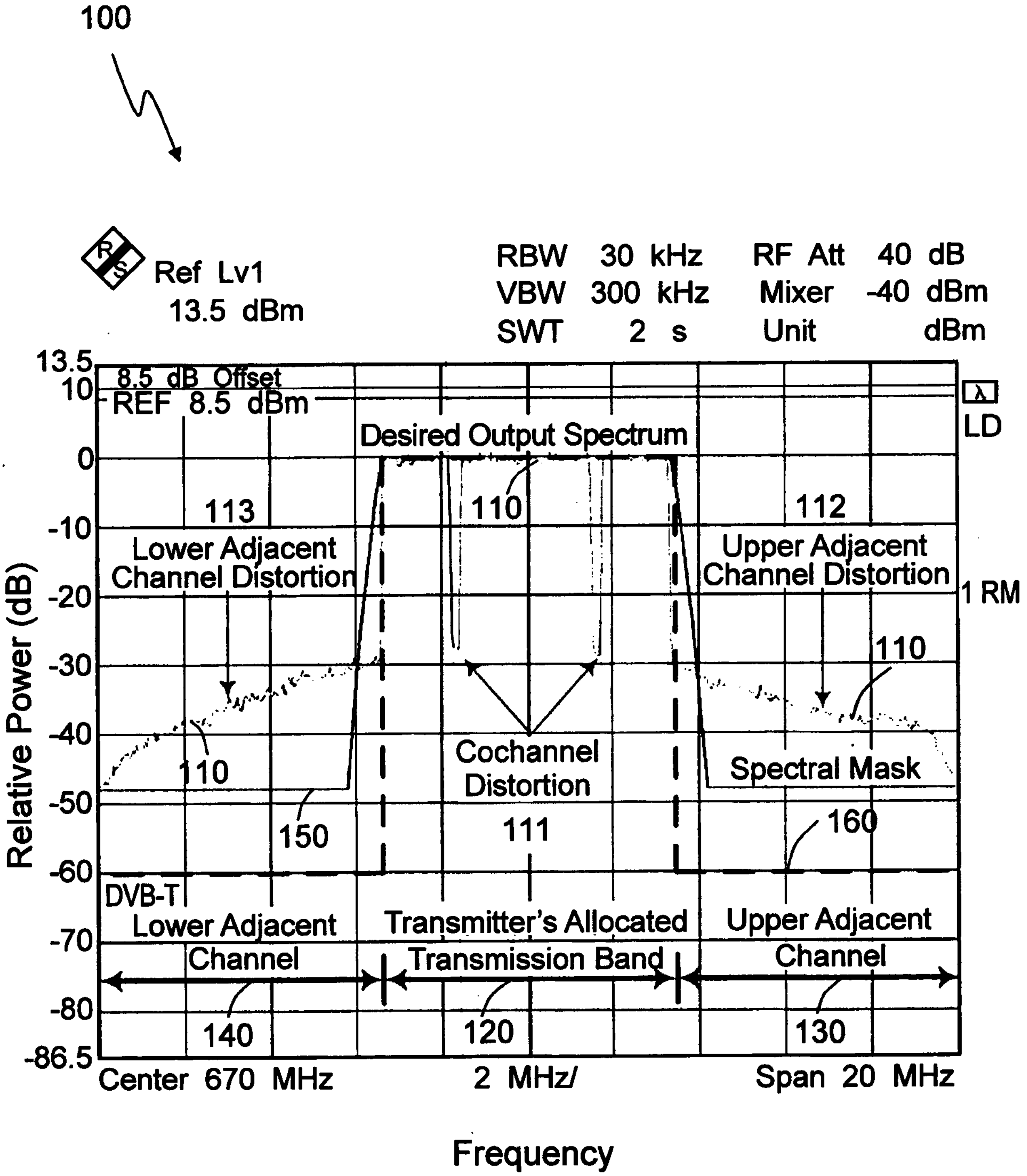

FIG 1
(PRIOR ART)

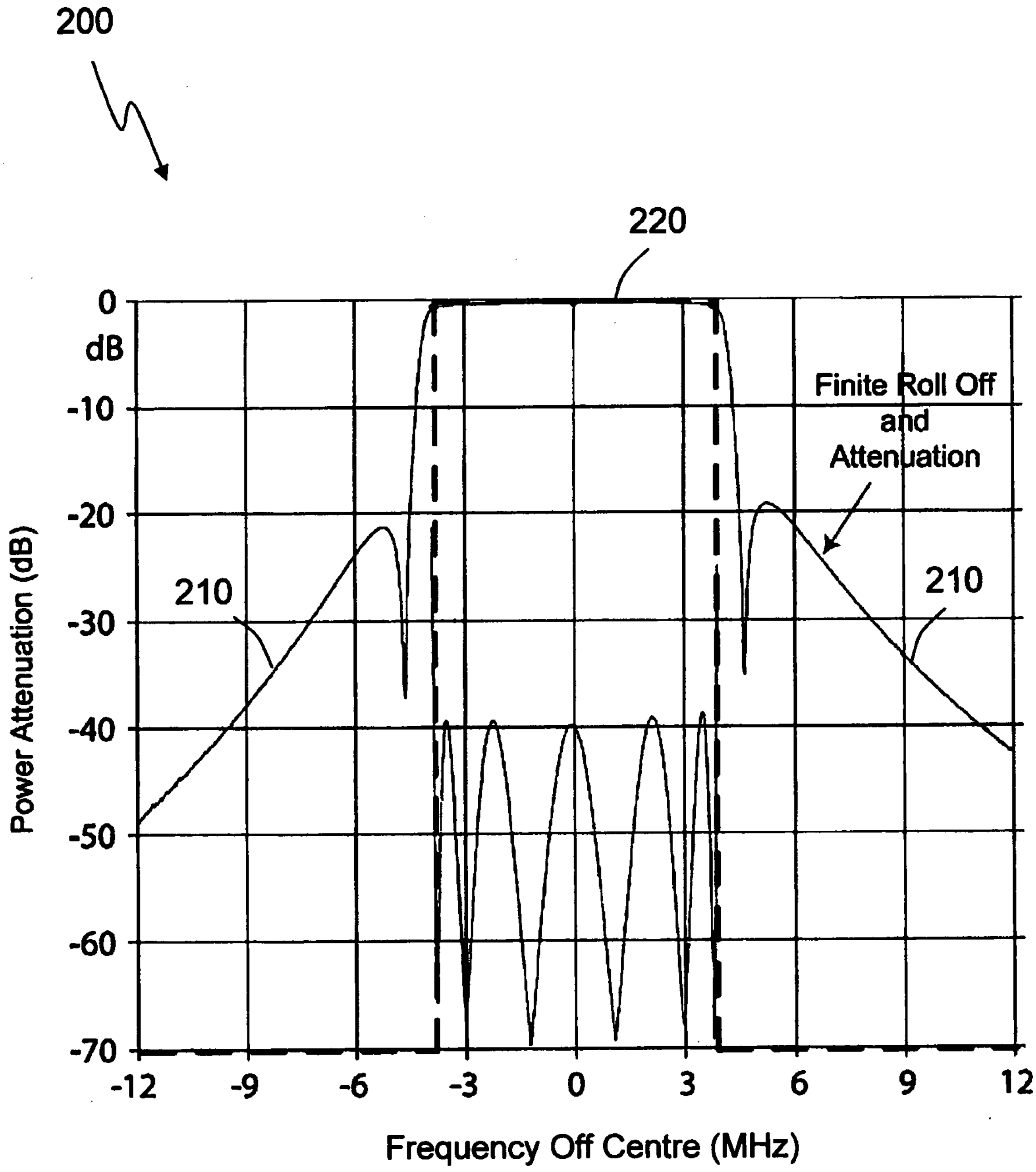

FIG 2
(PRIOR ART)

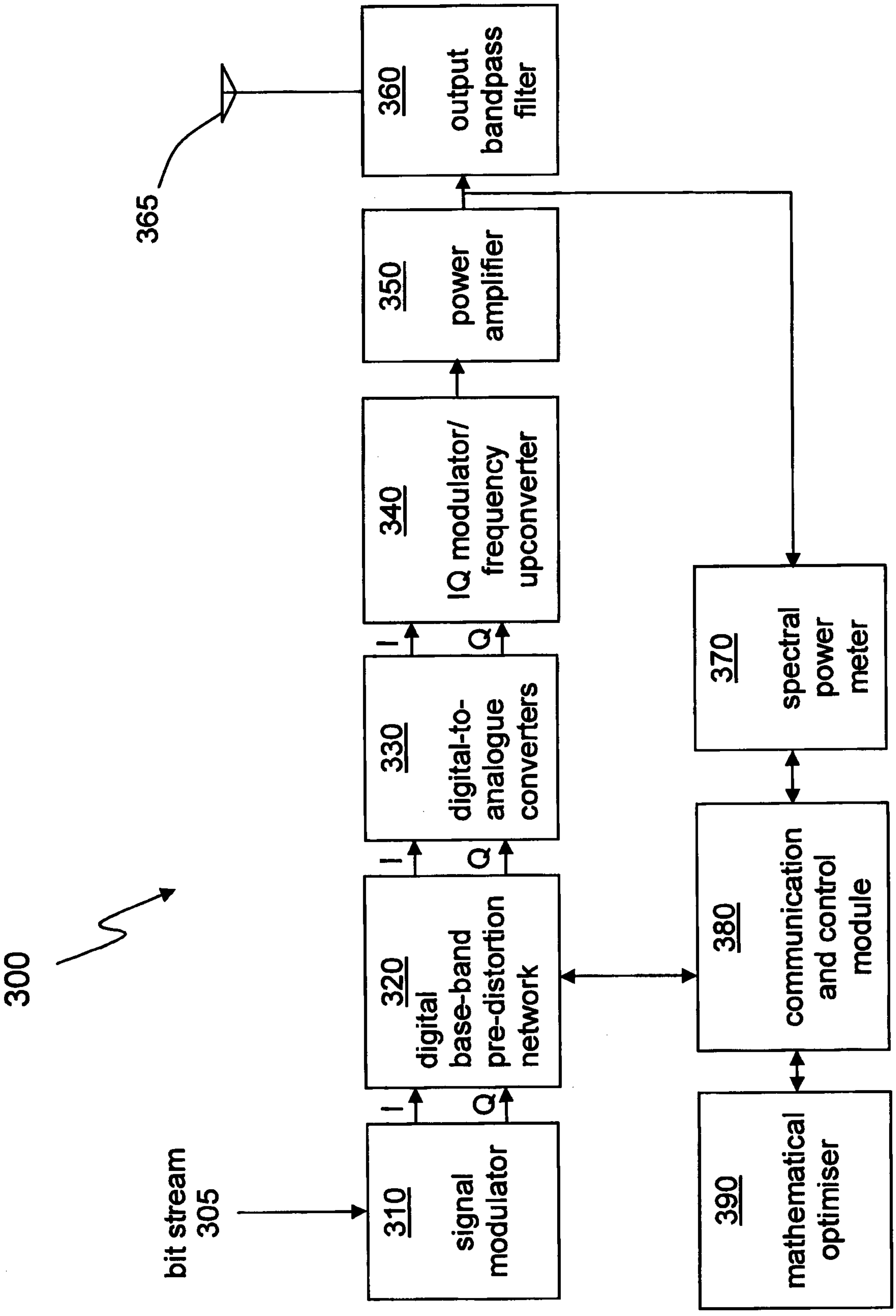
300
bit stream 305
310 signal modulator
I
Q
320 digital base-band pre-distortion network
I
Q
330 digital-to-analogue converters
I
Q
340 IQ modulator/ frequency upconverter
350 power amplifier
360 output bandpass filter
365
370 spectral power meter
380 communication and control module
390 mathematical optimiser

FIG 3

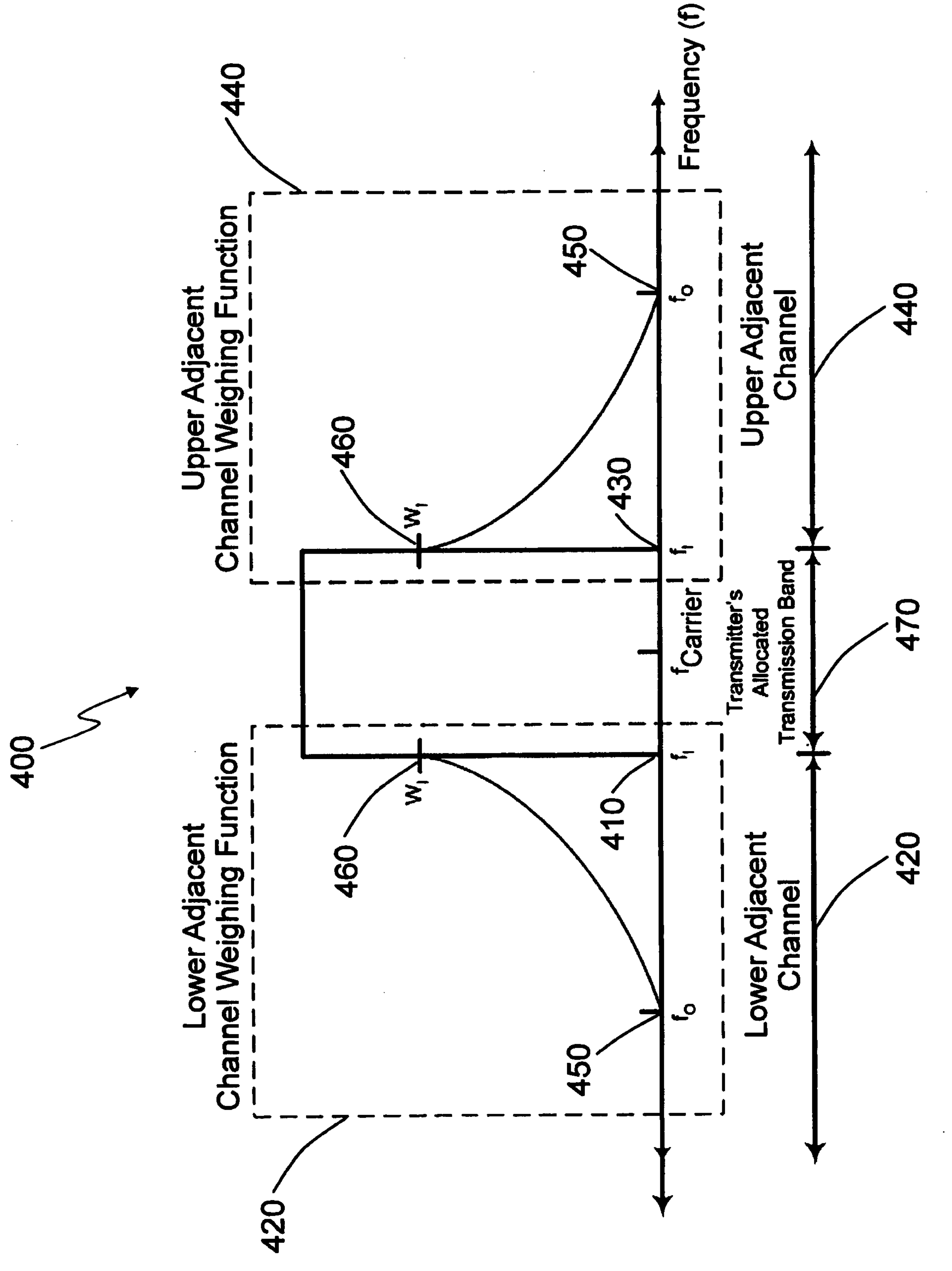
400
Frequency (f)
Upper Adjacent Channel Weighing Function
Lower Adjacent Channel Weighing Function
440
420
450
460
430
410
w1
f1
f0
fCarrier
Transmitter's Allocated Transmission Band
470
Upper Adjacent Channel
Lower Adjacent Channel

FIG 4

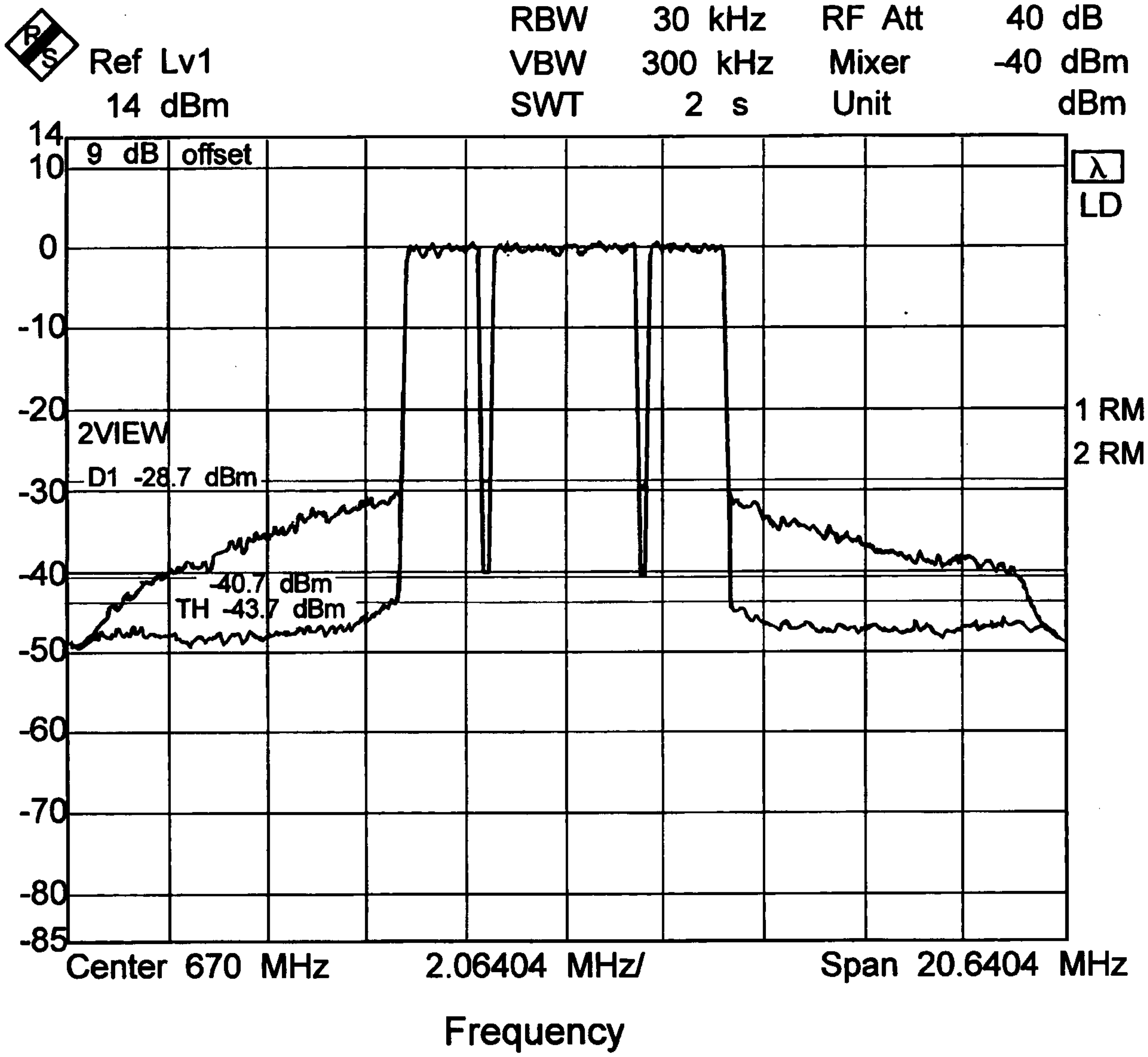
Ref Lv1
14 dBm
RBW 30 kHz
VBW 300 kHz
SWT 2 s
RF Att 40 dB
Mixer -40 dBm
Unit dBm
9 dB offset
λ
LD
1 RM
2 RM
2VIEW
D1 -28.7 dBm
-40.7 dBm
TH -43.7 dBm
14
10
0
-10
-20
-30
-40
-50
-60
-70
-80
-85
Center 670 MHz
2.06404 MHz/
Span 20.6404 MHz
Frequency

FIG 5

INTERNATIONAL SEARCH REPORT

International application No.
PCT/AU2011/001690

A. CLASSIFICATION OF SUBJECT MATTER

Int. Cl.

H04L 27/00 (2006.01)

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
DWPI & EPODOC - Keywords (multi carrier, CDMA, RF, baseband, predistortion) and like terms.

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 2010/0027690 A1 (LIANG et al) 4 February 2010 Abstract, paragraphs [0018]-[0019]	
A	US 2003/0133404 A1 (CASTELAIN et al.) 17 July 2003 Abstract, paragraphs [0004]-[0008], [0016]-[0018]	

☐ Further documents are listed in the continuation of Box C ☒ See patent family annex

*	Special categories of cited documents:		
"A"	document defining the general state of the art which is not considered to be of particular relevance	"T"	later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"E"	earlier application or patent but published on or after the international filing date	"X"	document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"L"	document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)	"Y"	document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"O"	document referring to an oral disclosure, use, exhibition or other means	"&"	document member of the same patent family
"P"	document published prior to the international filing date but later than the priority date claimed		

Date of the actual completion of the international search 7 February 2012	Date of mailing of the international search report 10 February 2012
Name and mailing address of the ISA/AU AUSTRALIAN PATENT OFFICE PO BOX 200, WODEN ACT 2606, AUSTRALIA E-mail address: pct@ipaustralia.gov.au Facsimile No. +61 2 6283 7999	Authorized officer **JAE CHO** AUSTRALIAN PATENT OFFICE (ISO 9001 Quality Certified Service) Telephone No : +61 2 6222 3668

This Annex lists the known "A" publication level patent family members relating to the patent documents cited in the above-mentioned international search report. The Australian Patent Office is in no way liable for these particulars which are merely given for the purpose of information.

Patent Document Cited in Search Report		Patent Family Member					
US	2010027690	CN	101640660	TW	201012143		
US	2003133404	EP	1328083	JP	2003249911	US	7315530

Due to data integration issues this family listing may not include 10 digit Australian applications filed since May 2001.

END OF ANNEX