MODE ANALYSER

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I. Circuit Description

A. Functional Description

The Mode Analyser is a double tracking spectrum analyser with two analog outputs indicating the amplitude of the spectral lines falling within each of the two bandpass frequencies. These outputs can be controlled to operate in either a logarithmic or linear mode. The center frequencies of the two bandpass filters are controlled by the frequencies of two pairs of sinusoidal signals in quadrature (90° relative phase shift), a high frequency pair with frequency $f_H$ and a low frequency pair with frequency $20$ kHz + $f_S$. The two center frequencies are $f_H - f_S$ and $f_H + f_S$, and the bandwidth of the two filters is 700 Hz. The high frequency quadrature signals are generated by Quadrature Revolution Frequency Generator MPS/BR 236.11, and the low frequency quadrature signals are generated by Quadrature VCO MPS/BR 236.14.

This function is realized as follows, Fig. 1. The input signal is mixed with the high frequency quadrature signals in two mixers. After low-pass filtering the two signals only contain the difference frequency mixing products of frequency $f_S$ and are mixed in the second pair of mixers with the low frequency quadrature signals of frequency $f_L$. Sum ($f_L + f_S$) and difference ($f_L - f_S$) frequency mixing products appear on the outputs of the mixers. The sum of the outputs $V_7$ of these two mixers will contain only the difference component $f_L - f_S$ if the input signal is a lower side-band with frequency $f_H - f_S$, and only the sum component $f_L + f_S$ if the input signal is an upper side-band with frequency $f_H + f_S$. The difference of the outputs $V_7^*$ of the two mixers will contain only the difference component $f_L - f_S$ if the input signal is an upper side-band, and only the sum component $f_L + f_S$ if the input signal is a lower side-band. See Fig. 2. $f_L$ is chosen to be $f_L = f_C + f_S$, so only the difference frequency component $f_L - f_S = f_C$ will pass through the fixed bandpass filter with center frequency $f_C = 20$ kHz. An input signal with frequency $f_H - f_S$ will thus only give a signal at $V_7$, whereas an input signal with frequency $f_H + f_L$ only gives a signal at $V_7^*$. 
Fig. 1  MODE ANALYZER, BLOCK DIAGRAM
Fig. 2 MODE ANALYZER, MIXING PROCESSES
An input signal with frequency \( f_H + 2f_C + f_S \) does in principle also produce a signal \( V_7^* \) with frequency \( f_C \), but this is strongly attenuated by the low-pass filter after the first mixer pair.

The outputs of the two 20 kHz bandpass filters are converted to DC voltages by linear amplitude detectors, either directly (linear mode) through logarithmic converters (logarithmic mode).

The outputs of the two bandpass filters are externally available. They can be used together with the low frequency quadrature signals to restore the side-band frequency \( f_S \) in an SSB mixer. (Used in RF-knockout).

B. Circuit Details

See circuit diagram MPS/BR 236.19.1001. The low-pass filters \( H_1(s) \) in Fig. 1) are third order Chebyshev 1 dB ripple filters with \( f_{3dB} = 15 \) kHz and a Bode cut-off \( f_0 = 11 \) kHz. The poles are:

\[
p = \begin{cases} 
(-0.313 \pm j 1.224) & 2\pi*11*10^3 \\
(-0.626) & 2\pi*11*10^3 
\end{cases}
\]

The two complex poles are realized by an active RC filter (R6/R7-C7/C8) R15/R18 - R16/R17-C9/C10-IC21, which provides, at the same time common mode rejection. The real pole is realized by an RC filter (R19/C13). Precision (1%) components are used to ensure tracking of phase and amplitude of the two filters.

This filter cuts off steeper than a Butterworth filter of the same order and the same 3 dB frequency. The frequencies \( f_L + f_C = 2f_C + f_S = 41 \) to 55 KHz, which give undesired mixing products with frequency \( f_C \) in the second mixer pair are attenuated 35 to 42 dB. The filters include a high-pass cutting off below 190 Hz (R6/C5 - R7/C6). This provides a complete suppression of input signal components at \( f_H^* \). (Unequal bunch lines).
The second pair of mixers have offset adjustments of one of their inputs to lower the feedthrough component from the $f_L$ (FS) signal.

All input terminals to the MC1495 mixers are equipped with 200 ohm parasite dampers, as the negative real part of the input impedance at certain frequencies (~150 MHz) can be as big as 150 $\Omega$.

The bandpass filters are third order 0.5 dB ripple Chebyshev filters with a center frequency $f_c = 20$ kHz, a half bandwidth $\Delta f_{3dB} = 350$ Hz, and a Bode cut-off at $\Delta f_o = 270$ Hz. The bandwidth is chosen as a compromise between good rejection of undesired frequencies and fast response time ($T_r = 0.5$ ms). The poles required to realize this are:

$$p = \left\{ \begin{array}{l} +j \ 2\pi*20K - 0.70*2\pi*270 \\ +j \ 2\pi*20K + j \ 1.14*2\pi*270 - 0.35*2\pi*270 \end{array} \right.$$ 

The filter is realized by three LC circuits in a T-network in the feedback path of an operational amplifier (IC4 - IC5), which also serves as summing point for the outputs of the mixers. Beyond the cut-off frequency, the filter rolls off with -18 dB/oct until the impedance of the series LC circuit becomes equal to the parallel impedance of the two parallel LC circuits: $|Z_s| = |Z_p||Z_p|$. At this frequency a pair of complex zeros causes a dip in the response and the further roll-off continues with -6 dB/oct. The zeros are located at:

$$Z = +j \ 2\pi*20K + j \ 4.8*2\pi*270.$$ 

The zeros improve considerably the damping of frequencies 1000 to 1500 Hz from the center frequency. The damping is > 49 dB for $|\Delta f| > 1200$ Hz.

The large phase shift variations of the feedback network makes the operational amplifier (IC4 - IC5) only conditionally stable. Diodes D1 - D2 prevent latch-up in an oscillating state, where saturation would lower the effective loop gain.
The air gap (or $\mu'$) of the three ferrite potcores L1, L2, and L3 has been selected to give a temperature coefficient for the inductance which cancels the temperature coefficient for the precision polystyrene capacitors C20, C23, and C24.

The logarithmic converter is made by three 30 dB sections of a logarithmic amplifier (SN 76502) ensuring more than 60 dB logarithmic range. Series resistor R50 prevents the input voltage range of the medium level section (pin 7) to be exceeded, whereas clamp diodes (D5 - D6 - D7 - D8) does it for the low level section (pin 9).

The amplitude detector is a linear full-wave average detector build around IC15 - IC16 - IC17. The cut-off frequency of the second order roll-off in the feedback path of IC16 - IC17 is 1200 Hz, which ensures sufficient damping (57 dB) of the $2f_C = 40$ kHz ripple frequency component without increasing the response time imposed by the bandwidth of the bandpass filter.

C. Signal Equations

The input signal:

$$V_1(t) = a_1 \cos \omega t = a_1 \cos(\omega_H + \omega_S)t$$

$$= \Re \{a_1 e^{j(\omega_H + \omega_S)t}\}$$

is mixed with the high frequency quadrature pair:

$$V_2(t) = a_2 \cos \omega_H t = \frac{a_2}{2} e^{j\omega_H t} + \frac{a_2}{2} e^{-j\omega_H t}$$

$$V_2^*(t) = a_2 \cos(\omega_H t - \frac{\pi}{2}) = \frac{a_2}{2} e^{j(\omega_H t - \frac{\pi}{2})} + \frac{a_2}{2} e^{-j(\omega_H t - \frac{\pi}{2})}$$

producing the mixer outputs:
\[ V_3(t) = V_1^*V_2 = \text{Re} \left\{ \frac{a_1a_2}{2} e^{j(2\omega_H + \omega_S)t} + \frac{a_1a_2}{2} e^{j\omega_S t} \right\} \]

\[ V_3^*(t) = V_1V_2^* = \text{Re} \left\{ \frac{a_1a_2}{2} e^{j((2\omega_H + \omega_S)t - \frac{\pi}{2})} + \frac{a_1a_2}{2} e^{j(\omega_S t + \frac{\pi}{2})} \right\} \]

which after low-pass filtering becomes:

\[ V_u(t) = \text{Re} \left\{ \frac{a_1a_2}{2} H_1(j\omega_S) e^{j\omega_S t} \right\} \]

\[ V_u^*(t) = \text{Re} \left\{ \frac{a_1a_2}{2} H_1(j\omega_S) e^{j(\omega_S t + \frac{\pi}{2})} \right\} \]

which are mixed with the low frequency quadrature pair:

\[ V_5 = a_5 \cos \omega_L t = \frac{a_5}{2} e^{j\omega_L t} + \frac{a_5}{2} e^{-j\omega_L t} \]

\[ V_5^* = a_5 \cos(\omega_L t - \frac{\pi}{2}) = \frac{a_5}{2} e^{j(\omega_L t - \frac{\pi}{2})} + \frac{a_5}{2} e^{-j(\omega_L t - \frac{\pi}{2})} \]

producing the mixer outputs:

\[ V_6 = V_u^*V_5 = \text{Re} \left\{ \frac{a_1a_2a_5}{4} H_1(j\omega_S) e^{j(\omega_S + \omega_L)t} \right\} + \frac{a_1a_2a_5}{4} H_1(j\omega_S) e^{j(\omega_S - \omega_L)t} \]

\[ V_6^* = V_uV_5^* = \text{Re} \left\{ \frac{a_1a_2a_5}{4} H_1(j\omega_S) e^{j(\omega_S + \omega_L)t} \right\} - \frac{a_1a_2a_5}{4} H_1(j\omega_S) e^{j(\omega_S - \omega_L)t} \]
which are added and subtracted:

\[ V_7 = V_6 + V_6^* = \text{Re}\left\{ \frac{a_1 a_2 a_5}{2} H_1(j\omega_S) e^{j(\omega_S + \omega_L) t} \right\} \]

\[ V_7^* = V_6 - V_6^* = \text{Re}\left\{ \frac{a_1 a_2 a_5}{2} H_1(j\omega_S) e^{j(\omega_S - \omega_L) t} \right\} \]

which after bandpass filtering becomes:

\[ V_8 = \text{Re}\left\{ \frac{a_1 a_2 a_5}{2} H_1(j\omega_S) H_2(j(\omega_S + \omega_L)) e^{j(\omega_S + \omega_L) t} \right\} \]

\[ V_8^* = \text{Re}\left\{ \frac{a_1 a_2 a_5}{2} H_1(j\omega_S) H_2((j(\omega_S - \omega_L)) e^{j(\omega_S - \omega_L) t} \right\} \]

The low frequency quadrature pair has the frequency:

\[ \omega_L = \omega_C + \omega_{SO} \]

The \( V_8 \) signal will be within passband if:

\[ \omega_C - \frac{\Delta\omega_C}{2} < |\omega_S + \omega_L| < \omega_C + \frac{\Delta\omega_C}{2} \]

which correspond to input frequencies around:

\[ \omega = \omega_H - \omega_{SO} \quad \text{(desired lower side-band)} \]

and:

\[ \omega = \omega_H - (2\omega_C + \omega_{SO}) \]
which is an undesired passband, which is strongly suppressed because:

$$|H_1(j(2\omega_C + \omega_{SO}))| << |H_1(j\omega_{SO})|$$

The $V_8^*$ signal will be within passband if:

$$\omega_C - \frac{\Delta\omega_C}{2} < |\omega_S - \omega_L| < \omega_C + \frac{\Delta\omega_C}{2}$$

which correspond to input frequencies around:

$$\omega = \omega_H + \omega_{SO}$$  (desired upper side-band)

and:

$$\omega = \omega_H + (2\omega_C + \omega_{SO})$$

which again is suppressed by $H_1(j\omega)$.

If we had chosen $\omega_L = \omega_C - \omega_{SO}$, we would also get bandpass for $\omega = \omega_H + \omega_{SO}$, but the undesired bandpass would be at $\omega = \omega_H + (2\omega_C - \omega_{SO})$, which would be much harder to suppress by $H_1(j\omega)$. 
II. Characteristics

a) Inputs

FR, FR* : Input level : 0.4 Vp (2.04 dBm)
Input impedance : 50 Ω
Frequency range, \( f_H \) : 0.6 – 15 MHz
DC coupled.

FS, FS* : Input level : 2 Vp
Input impedance : > 1 MΩ
Frequency range, \( f_L \) : 21 kHz – 35 kHz
DC coupled.

IN : Input impedance : 50 Ω
Can be changed to 100 Ω if 2 units are to be used in parallel
AC coupled.
Saturation limit, 1. mixer:
\[ |f_{IN} - f_H| > 30 \text{ kHz} \text{ or } |f_{IN} - f_H| < 10 \text{ Hz} \]
\( V_{IN,\text{MAX}} = 0.5 \text{ Vp (}+4 \text{ dBm)} \)

Saturation limit, 2. mixer or bandpass filter:
\[ 200 \text{ Hz} < |f_{IN} - f_H| < 15 \text{ kHz} \]
\( V_{IN,\text{MAX}} = 22.5 \text{ mVp (}-23 \text{ dBm)} \)

LIN/LOG : CMOS/TTL logic input.
10 kΩ pull-up to +6 V.
0 V : Logarithmic mode.
+6 V: Linear mode.

b) Outputs

FO+, FO- : 20 kHz bandpass output.
Max. output level: 9 Vp unterminated.
4.5 Vp (+23 dBm) terminated in 50 Ω
Output impedance : 50Ω
DC coupled.

Gain from IN input : 400X (52 dB) unterminated.
200X (46 dB) terminated.

\textbf{OUT+, OUT-} : Detector DC output.

Output impedance : 50 Ω
DC coupled.

Max. output voltage (Linear mode): $+9 V_{\text{DC}}$ unterminated.

Sensitivity (linear mode) : $+0.4 V_{\text{DC}}/\text{mVp}$ (unterminated)

Sensitivity (logarithmic mode) : $0.1 V_{\text{DC}}/\text{dB}$. (unterminated)
$+6.00 V_{\text{DC}}$ (unterminated) for
$-28 \text{ dBm} (12.5 \text{ mVp})$ input level.

$0.00 V_{\text{DC}}$ (unterminated) for
$-88 \text{ dBm} (12.5 \mu\text{Vp})$ input level. Relative error in
logarithmic mode over 65 dB range $-88 \text{ dBm}$ to $-23 \text{ dBm}:
< ± 30 \text{ mV}_{\text{DC}} = ± 0.3 \text{ dB}.$

\textit{c) Other}

Bandpass frequency range : $f = f_H + f_S$;

1 kHz $\leq f_S \leq 15$ kHz.

$f_S = f_L - 20$ kHz.

Change in sensitivity over $f_S = 1$ to 14 kHZ range:

$+0/-1$ dB.

Bandpass bandwidth : $-1$ dB : 600 Hz (+ 300 Hz)
$-3$ dB : 700 Hz (+ 350 Hz)
$-49$ dB : for $|\Delta f| > 1200$ Hz.

Undesired bandpass at \textbf{OUT+} : $f_H - f_S$, $f_H + 2f_C + f_S$
\textbf{OUT-} : $f_H + f_S$, $f_H - (2f_C + f_S)$
suppressed 35 - 40 dB

Response time : \sim 0.5 ms.
III. Calibration Procedure

1) Check proper orientation of all IC’s before putting power on the module. Check that input impedance is 50 Ω.

2) Remove jumpers IC5/pin 4 - R39/R40/R50 and IC 105/pin 4 - R139/R140/R150.

3) Verify that 'LIN' is lit. Adjust 'DET. INP. OFFS.-' (P10) for 0 $V_{DC}$ (< ± 100 mV) output on IC 15/pin 10.

4) Adjust 'DET. OUTP. OFFSET-' for 0 $V_{DC}$ (< ± 5 mV) output on OUT-.

4a) Repeat 3) and 4) until both correct.

5) Connect Network Analyser (hp 3570A/3320B) output A to FO-. Set LEVELLING to ON and frequency to 20.00 kHz.

6) Adjust Network Analyser output level (approximately +19.0 dBm) until OUT- = +5.00 $V_{DC}$.

7) Change to 'LOG' mode by inserting a 50 Ω termination into LIN/LOG.

8) Turn 'HIGH LEV. GAIN-' more than 5 turns clockwise and then 3 turns counter-clockwise.

9) Adjust 'LOG. SLOPE-' for OUT- = +6.00 $V_{DC}$.

10) Reduce Network Analyser output level by 30 dB, and adjust 'LOG OUT OFFS.-' for OUT = + 3.00 $V_{DC}$.

11) Increase Network Analyser level to previous value and repeat 9) and 10) until both correct.

12) Repeat 3) to 11) for '+' channel.

13) Connect jumpers removed in 2).
14) Prepare the test set-up on Fig. 3. (A-FR) and (A*-FR*) cables must be of equal length. Set signal generator to 1 Vpp, 5.000 MHz. Set power supply for 22.0 kHz on Quadrature VCO. Set spectrum analyser as follows:

Center frequency : 5.998 MHz
Scan width : 100 Hz/div
Scan time : 50 ms/div
Scan mode : Internal
Scan trigger : Automatic
Tuning stabiliser : On

and tracking generator as follows:

Output level : -60 dBm
(Function : Track Analyser)

Adjust scope horizontally to have the sweep covering the whole screen (10 div.). Use 0.2 V/div. vertical sensitivity (2 dB/div.).

15) Set Network Analyser to -44 dBm, 20.000 KHz and connect its output to the test input on the Mode Analyser printed circuit board (R32).

16) Adjust Spectrum Analyser center frequency to center the beating pattern on the oscilloscope screen. Repeat this check from time to time during the following. Remove Network Analyser signal.

17) Adjust \( f_h \) \((L1)\), \( f_{L} \) \((L2)\), and \( f_{C} \) \((L3)\) to obtain a symmetric, centered, and approximately 700 Hz wide (-3 dB) bandpass response.

18) Set Network Analyser to Spectrum Analyser center frequency. (Approximately 5.998 MHz), level = 28.00 dBm. Switch to 'LIN' mode by removing 50 Ω termination. Connect Network Analyser output to Mode Analyser IN. Adjust Network Analyser frequency to obtain 20.00 kHz ± 20 Hz at FO-output. Measure the OUT- voltage and calculate how many dB the bandpass filter gain will have to be changed to obtain +5.0 V_{DC} at OUT-. 
Fig. 3 MODE ANALYZER, TEST SET-UP

- NETWORK ANALYZER hp 3590A
- FREQUENCY SYNTHESIZER hp 3520 B
- PEN LIFT
- OSCILLOSCOPE hp Tektronix
- DVM
- TRACKING GENERATOR hp 8434 A
- SPECTRUM ANALYZER hp 8533 B
- MODE ANALYZER MPS/BR 236.19
- QUADRUATURE V.C.O. MPS/BR 236.14
- COUNTER hp 5302 A
- SIGNAL GENERATOR TEKTRONIX TYPE 191
- POWER SUPPLY 0 - +4.5 V D.C.

IN - 5000 MHz
IN - 5.98 MHz
IN - 6.002 MHz
OUT - 22 KHz
OUT - X
OUT - Y
OUT - Z
OUT - A
OUT - FR
OUT - FS
OUT - +2.2 V D.C.
19) Reconnect tracking generator to Mode Analyser input. Switch to
   LOG mode. Adjust $f_h$ (L1) and $f_L$ (L2) to obtain the required gain
   change. Maintain symmetry and centering. Readjust eventually
   $f_c$ (L3).

20) Repeat 18) and 19) until no further correction is required. Secure
   L1, L2, and L3 screws with hot wax. Let coils cool off. Check 18)
   and 19).

21) Repeat 14) to 20) for '+' channel. Center frequency 6.002 MHz.

22) Switch to 'LIN' mode. Remove input signal. Adjust Power Supply for
   a Quadrature VCO frequency of 20.0 kHz.

23) Adjust 'OFFSET' and 'OFFSET'' to obtain minimum output of OUT+ and
   OUT− (< 30 mV<sub>DC</sub>).

24) Set Quadrature VCO frequency back to 22.00 kHz.

25) Connect Network Analyser through 20 dB attenuator to Mode Analyser input.
   Set frequency to 5.998 MHz and level to −8 dBm. Adjust frequency to
   obtain 20.00 kHz ± 20 Hz at FO−. Check OUT− = 5.00 ± 0.2 V<sub>DC</sub> in 'LIN'
   mode.

26) Switch to 'LOG' mode. Adjust 'LOG. SLOPE-' to obtain OUT− = +6.00 V<sub>DC</sub>
   for Network Analyser level +8 dBm. (−28 dBm at input).

27) Set Network Analyser level to −33 dBm. (−53 dBm at input). Adjust
   'LOG. OUT OFFSET-' to obtain OUT− = +3.50 V<sub>DC</sub>.

28) Repeat 26) and 27) until both correct.

29) Set Network Analyser level to −64 dBm (−84 dBm at input). Adjust 'LOG.
   INF. OFFS.-' to obtain OUT− = +0.40 V<sub>DC</sub>.

30) Set Network Analyser level to −58 dBm (−78 dBm at input). Adjust
   'LOW LEV.-' to obtain OUT− = +1.00 V<sub>DC</sub>.

31) Repeat 29) and 30) until both correct.
32) Repeat 26), 27), 29) and 30 until all correct.

33) Set Network Analyser level to -18 dBm (-38 dBm at input). Check OUT- = 5.00 V DC. If more than 0.02 V too big, set Network Analyser to -8 dBm and adjust 'HIGH LEV. GAIN' to obtain a 0.1 V increase in OUT- voltage (6.1 V). If the value was more than 0.02 V too small, adjust 'HIGH LEV. GAIN' to obtain a 0.1 V decrease in OUT- voltage (5.9 V).

34) Repeat 26) to 33) until all correct.

35) Repeat 25) to 34) for '+ ' channel. Center frequency 6.002 MHz.

36) Connect tracking generator to Mode Analyser input again, level -28 dBm. Set Spectrum Analyser to:

   Center frequency : 6.000 MHz
   Scan width : 1 kHz/div.
   Scan time : 50 ms/div.

Connect both OUT- and OUT+ to Y channels on scope, 1 V/div or 10 dB/div.

37) Adjust 'BAL-' to get maximum suppression (∼ 40 dB) of upper side-band on OUT- trace.

38) Adjust 'BAL+' to get maximum suppression (∼ 40 dB) of lower side-band on OUT+ trace.

39) Remove R1 = 100 Ω if 100 Ω input impedance is required.

40) Label the module, that it has been checked and adjusted.

Distribution:

BR Electronics section
and as requested to Miss M. Innocenti