NEGATIVE GROUP DELAY SYNTHESISER
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Indexing terms: Network synthesis, Integrated circuits, Microwave circuits

A network that synthesises negative group delay is presented. The conditions required for achieving negative group delay, with a simple R-L-C tuned circuit, are briefly introduced. A hybrid MIC realisation of the network, operating at 1 GHz, is described and its measurements are reported.

Introduction: The group delay \( \tau \) of an analogue network can be defined by the following expression:

\[
\tau = -\frac{\phi}{\omega}
\]

where \( \phi \) is the insertion phase of the network and \( \omega \) the angular frequency. This suggests that if the phase gradient is positive, the group delay will be negative. A positive phase gradient anomaly was recently identified [1] in simulations with a simple reflection-type phase shifter topology [2], having simple R-L-C series tuned circuit reflection terminations. The insertion phase of the network can be expressed as follows [1]:

\[
\phi = \frac{\angle Z_{in}}{2} \left( 1 - \frac{2\omega}{\omega_0} \right)
\]

where \( \angle Z_{in} \) is the angle of voltage reflection for the reflection termination and \( \omega_0 \) the angular frequency of maximum coupling. For the case of a simple R-L-C series tuned circuit reflection termination, it can be shown that [1]

\[
\angle Z_{in} = \tan^{-1} \left( -\frac{2\omega Z_L X_T}{R^2 - Z_L^2 + X_T^2} \right)
\]

and

\[
\frac{\partial \angle Z_{in}}{\partial \omega} = -\frac{2\omega Z_L X_T}{R^2 - Z_L^2 + X_T^2} \left( \frac{2X_T^2 - (R^2 - Z_L^2 + X_T^2)}{(2Z_L X_T)^2 + (R^2 - Z_L^2 + X_T^2)^2} \right)
\]

where \( Z_L \) is the input impedance of the coupled/direct ports (usually 50Ω), \( R \) the resistance, \( X_L \) the inductive reactance, \( X_C \) the capacitive reactance and \( X_T \) the total reactance (i.e. \( X_L + X_C \)).

At the angular frequency at which series resonance occurs, \( \omega_0 = 2\omega_0 \):

\[
\angle Z_{in} = \frac{4Z_L L}{Z_0^2 - R^2}
\]

therefore

\[
\tau = \frac{4Z_L L}{Z_0^2 - R^2} \times \frac{1}{2\omega_0}
\]

where \( L \) is the inductance and \( \omega_0 \) the frequency of maximum coupling. From this last expression, it can be deduced that as the resistance reduces from \( \omega_0 (Z_0^2 + 2Z_0 L f_c) \) to \( Z_0 \), the group delay will rapidly decrease from zero to minus infinity. In a similar way, as the resistance increases from zero to \( Z_0 \), the group delay will rapidly increase from \( 4L/Z_0 + 1/2\omega_0 \) to plus infinity.

MIC realisation: A synthesiser, operating at 1 GHz, was realised using microstrip techniques. The topology of this experimental synthesiser is illustrated in Fig. 1. A four-finger Lange coupler was employed and folded, to make the microwave integrated circuit (MIC) as compact as possible. The variable resistors were realised using cold FETs and the variable capacitors were implemented with chip varactor diodes. All the bias resistors and DC blocking capacitors were surface mounted. The combined inductance of the bond wires, which connect the capacitors and FETs together, form the inductors. A photograph of the experimental synthesiser is shown in Fig. 2. The active area has dimensions of only 1.1 x 1.3 cm².

Measurements: Conventional on-wafer probing techniques were adopted, to avoid any measurement degradation encountered with traditional microstrip launchers. To facilitate this, a coplanar waveguide-to-microstrip transition was developed, having low inductance wrap-around grounds. An HP8510B automatic network analyser and a Cascade Summit-9000 analytical probe station were used to perform the measurements.

Fig. 1 Topology of negative group delay synthesiser

Fig. 2 Photograph of compact 1 GHz synthesiser realisation

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The phenomenon of positive insertion phase gradient and the corresponding negative group delay frequency responses are shown in Fig. 3 and Fig. 4, respectively. The tuning curve of the synthesiser, at \( f_b = f_c = 1 \text{ GHz} \), is given in Fig. 5. From Figs. 4 and 5, it is apparent that as the gate bias potential increases, from \(-2.30 \text{ to } -1.8 \text{ V}\), the drain-source channel resistance decreases and the negative group delay rapidly increases. Also, as negative group delay increases, its bandwidth rapidly decreases. The bandwidth for the measured group delay at \(-10 \text{ ns}\) is approximately 40 MHz, whereas that for the group delay at \(-50 \text{ ns}\) is only 1 MHz. A group delay of \(-800 \text{ ns}\) was measured with a bandwidth of \(-75 \text{ kHz}\). The measured insertion loss performance of the synthesiser is shown in Fig. 6. Here, insertion loss rapidly increases as the group delay approaches either of the singularities, i.e. where \( R = Z_d \) and \( R = Z_c \). A high insertion loss can be overcome by embedding the synthesiser between two high gain amplifiers. One of the inherent advantages with the synthesiser is its excellent return loss performances. The measured input return loss was better than 26 dB across the whole bandwidth of interest and at all bias points. This is an important feature when the synthesiser is embedded between two high gain amplifiers, because the amplifiers are more likely to remain unconditionally stable if they are perfectly matched.

**Discussion:** When negative group delay is of the order of \( \mu \text{s} \), i.e. with a large value of resistance, the synthesiser could find applications in adaptive group delay equalisers. If tunable active inductors are employed in the reflection terminations, the characteristics of the group delay frequency response can be completely controlled. Alternatively, when the group delay is either in a sharp dip or on a sharp peak, the synthesiser can find applications in feedback oscillators which exhibit high levels of long term stability. The levels of carrier frequency drift will rapidly decrease as group delay approaches either of the singularities. Closer to singularities, the synthesiser could be used as an extremely narrowband tunable bandstop filter. All these applications are totally compatible with monolithic MIC technology; the latter two offer the additional advantage of not requiring off-chip high-Q resonators. With all the applications, however, there are penalties of high insertion loss and long noise figure.

It must be noted that the group delay of an analogue network is only equal to the transit time delay of a purely sinusoidal signal propagating through it when the network is lossless. Therefore, the difference in the levels of group delay and transit time delay increases as the insertion loss of the network increases.

**Conclusions:** The theoretical concepts for a practical negative group delay synthesiser have been introduced. A compact hybrid MIC synthesiser, operating at 1 GHz, was realised, the measured results of which have been reported and potential applications discussed. The synthesiser is totally compatible with monolithic MIC technology and if an active circulator is employed, instead of the coupler, a considerable reduction can be made in the active area required.

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PERFORMANCE OF FOUR-CHANNEL FDM CROSSCONNECT SWITCHING WITHOUT BIT LOSS

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Optical frequency switching without bit loss is demonstrated at the data rates of 156 and 622 Mbit/s in a four-channel frequency-division-multiplexing (FDM) crossconnect which employs fast tunable lasers and a fixed-tuned channel selection filter.

Optical frequency switching for crossconnect and packet switching applications is being investigated in order to significantly increase the transport and processing capabilities of optical networks [1-6]. From the viewpoint of the crossconnect system design, it is important to perform frequency switching without dropping any bits. The combined action of fast tunable lasers and fixed-tuned channel selection filters offers the possibility of realising an optical frequency-division-multiplexing (FDM) crossconnect that can achieve the needed switching performance. The important characteristics of rapidly-tunable lasers for frequency switching are switching time, dynamic tunability, and thermally-induced frequency drifts [1]. The switching time between channel frequencies should be less than the bit period which is determined by the data rate. For example, a switching time less than 1.6 ns is required to perform frequency switching without bit loss at 622 Mbit/s. The number of allowable destinations at the fixed channel spacing depends on the dynamic tunability. The packet holding time is strongly related to thermally-induced drift [4,7].

We report the successful testing of a four-channel FDM crossconnect that switches without bit loss at the data rates of 156 and 622 Mbit/s. The effect of the thermally-induced drift is also described.

Fig. 1 shows the schematic diagram of the experimental setup. A three-section DBR laser or DFB laser was used as the fast tunable transmitter. Switching between channel frequencies is performed by driving the respective DBR section and the centre section of the DBR and DFB lasers with a multilevel pulse pattern generator. The staircase signal is applied to the DBR tuning section or the DFB centre section. The optical data signal is amplitude shift keyed by an LiNbO3 modulator. A 10 GHz-spaced, four-channel Mach-Zehnder demultiplexer [8] in front of each receiver was used as a fixed-tuned channel selection filter. The demodulated signal outputs of the receivers were observed on a sampling oscilloscope.

A switching time of ~3 and 0.4 ns over the 60 GHz tunable frequency range was measured for the DBR and DFB lasers, respectively. This means that the DBR and DFB laser transmitters can realise optical frequency switching without bit loss at data rates less than 330 Mbit/s and 2.5 Gbit/s, respectively. The staircase waveform from the multilevel pulse generator has a 1 ns rise time, so the minimum switching time is limited by the risetime for the DFB laser transmitter. Therefore, frequency switching experiments were carried out at 156 and 622 Mbit/s with DBR and DFB laser transmitters, respectively. The demultiplexing characteristics of the 10 GHz-spaced, four-channel demultiplexer used as the fixed-frequency filter are shown in Fig. 2. The temperature of the demultiplexer was stabilised to within ±0.02°C. The crosstalk between the four channels was ~20 dB. Pseudorandom bit streams switched between four different destinations are shown in Fig. 3a and b for the DBR and DFB laser, respectively. The staircase waveforms corresponding to the bit streams are shown in the upper sections. A staircase waveform with almost equal steps could be applied to the DBR laser, while the steps were quite different for the DFB laser. The packet duration was at 30 and 13 ns for visualisation of the 8 bits per packet. Clearly visible is a 4 x 4 optical frequency switching without bit loss at 156 Mbit/s and at 622 Mbit/s. The experimental arrangement

Fig. 2 Demultiplexing characteristics of 10 GHz-spaced, four-channel demultiplexer used as fixed-tuned channel selection filter

Fig. 3 Pseudorandom bit stream switched between four different destinations using DBR laser transmitter at 156 Mbit/s and DFB laser transmitter at 622 Mbit/s.